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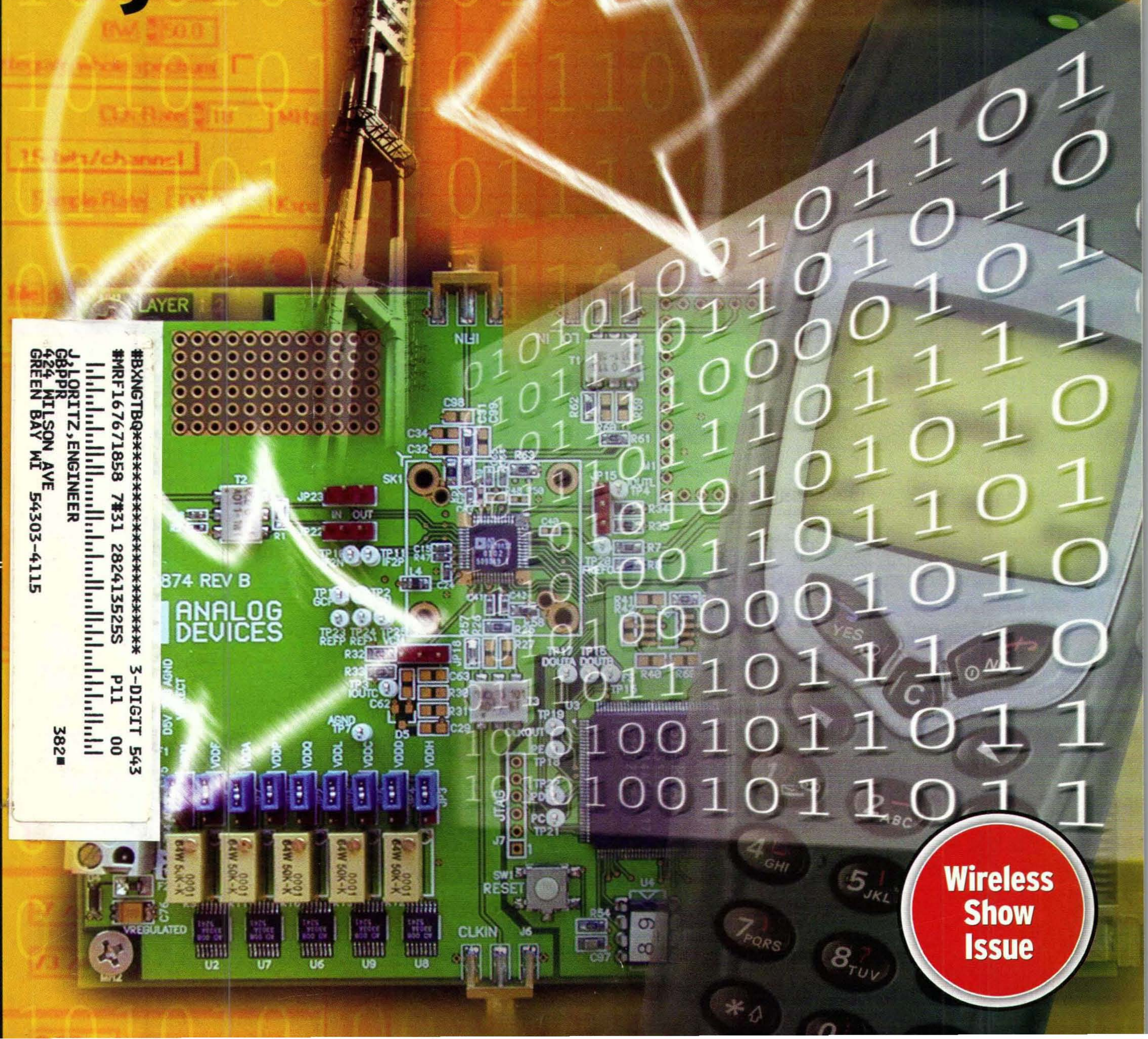
News
Previewing the Wireless Systems 2002 show

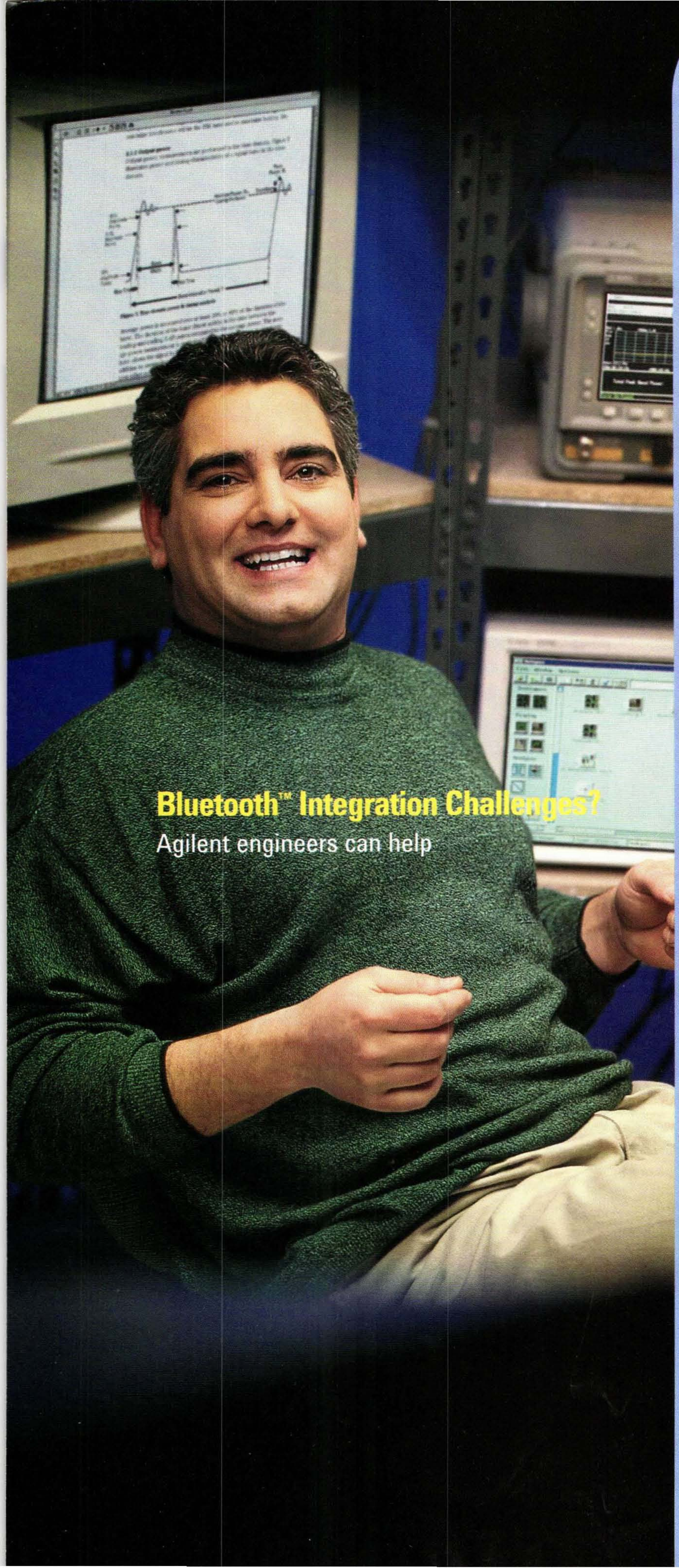
Design Feature
Gauge amplifier dynamic range

Product Technology
Top Products of 2001

Low-Power IF IC Digitizes 300 MHz



Wireless Show Issue



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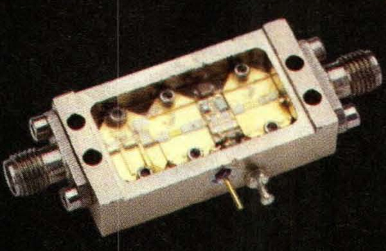

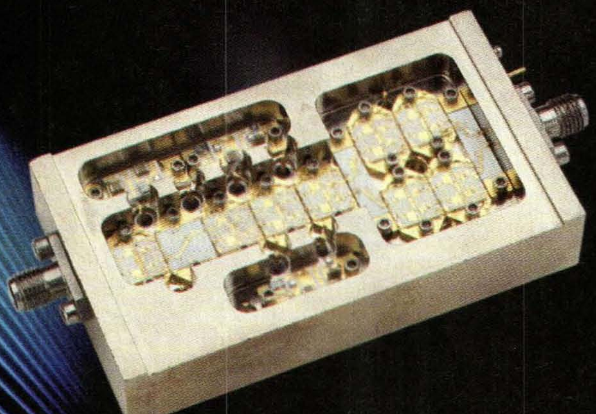

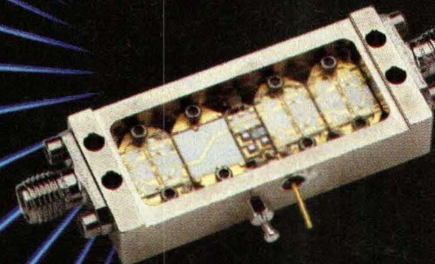

ULTRA BROAD BAND

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-203	0.5-18.0	20	5.0	2.5	7	17	2.0:1	250
JCA018-204	0.5-18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0-18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0-18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0-18.0	30	5.0	2.5	21	31	2.0:1	500

MULTI OCTAVE AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA04-403	0.5-4.0	27	5.0	1.5	17	27	2.0:1	550
JCA08-417	0.5-8.0	32	4.5	1.5	17	27	2.0:1	550
JCA28-305	2.0-8.0	22	5.0	1.0	20	30	2.0:1	550
JCA212-603	2.0-12.0	32	5.0	3.0	14	24	2.0:1	550
JCA618-406	6.0-18.0	20	6.0	2.0	25	35	2.0:1	600
JCA618-507	6.0-18.0	25	6.0	2.0	27	37	2.0:1	800

MEDIUM POWER AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35-1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1-3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9-6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0-12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0-18.0	22	4.0	2.0	25	35	2.0:1	700

LOW NOISE OCTAVE BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-3001	1.0-2.0	40	0.8	1.0	10	20	2.0:1	200
JCA24-3001	2.0-4.0	32	1.2	1.0	10	20	2.0:1	200
JCA48-3001	4.0-8.0	40	1.3	1.0	10	20	2.0:1	200
JCA812-3001	8.0-12.0	32	1.8	1.0	10	20	2.0:1	200
JCA1218-800	12.0-18.0	45	2.0	1.0	10	20	2.0:1	250

NARROW BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2-1.6	25	0.75	0.5	10	20	2.0:1	80
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1.0	0.5	10	20	2.0:1	90
JCA56-401	5.4-5.9	40	1.0	0.5	10	20	2.0:1	120
JCA78-300	7.25-7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0-9.5	25	1.2	0.5	13	23	1.5:1	150
JCA910-3001	9.5-10.0	25	1.2	0.5	13	23	1.5:1	150
JCA1112-3000	11.7-12.2	27	1.1	0.5	13	23	1.5:1	150
JCA1213-3001	12.2-12.7	25	1.1	0.5	10	20	2.0:1	200
JCA1415-3001	14.4-15.4	35	1.4	1.0	14	24	2.0:1	200
JCA1819-3001	18.1-18.6	25	1.8	0.5	10	20	2.0:1	200
JCA2021-3001	20.2-21.2	25	2.0	0.5	10	20	2.0:1	200

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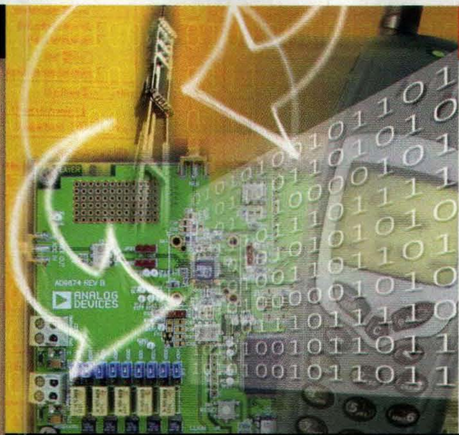

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Microwaves & RF (ISSN 0745-2993) is published monthly, except semi-monthly in December. Subscription rates for US are $80 for 1 year ($105 in Canada, $140 for International). Published by Penton Media, Inc., The Penton Building, 1300 E. 9th St., Cleveland, OH 44114-1503. Periodicals Postage Paid at Cleveland, OH and at additional mailing offices.

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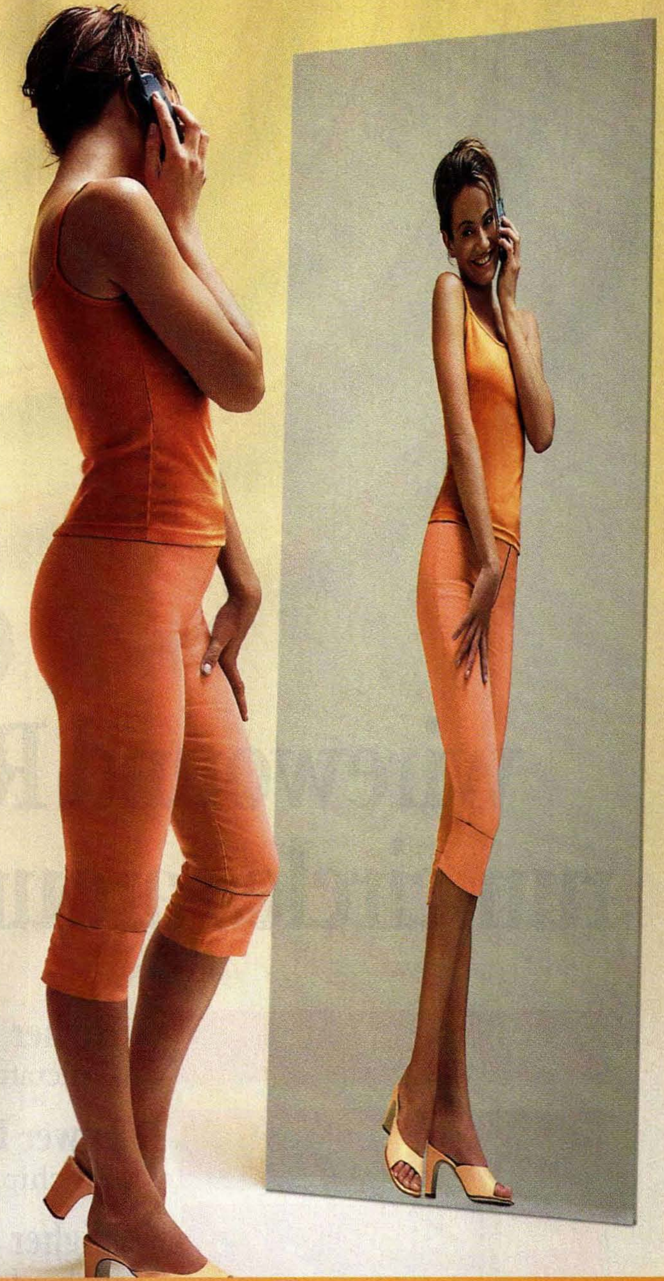

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TST0950	900-MHz LNA	GSM, ISM
TST0912	900-MHz PA	GSM
TST0951	1900-MHz SiGe LNA	DCS & PCS mobile phones
T7024	2.4-GHz SiGe Front End	ISM/Bluetooth
T0980	400/500-MHz SiGe Front End	Family radio (Walky Talky) & remote control applications

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LNA: Low Noise Amplifier

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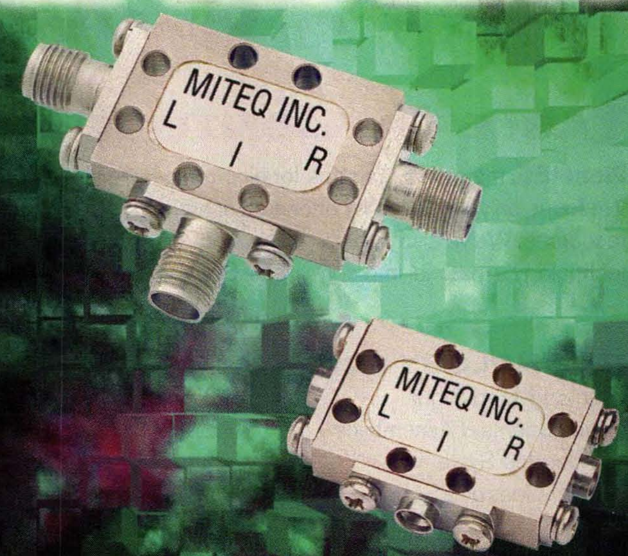

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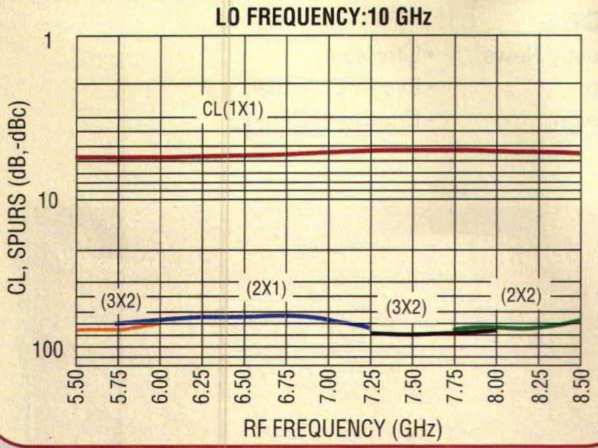
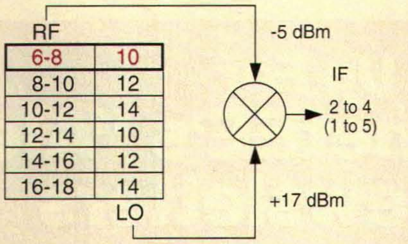


TYPICAL OPERATING BANDS



RF	LO
6-8	10
8-10	12
10-12	14
12-14	10
14-16	12
16-18	14

SPECIFICATIONS - Model TBR0618HA1/TBR0618HA1-S

RF/LO Input Frequency Range	6 to 18 GHz
IF Output Frequency Range	0.05 to 5 GHz
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Wrong Year

▶▶I WOULD LIKE to point out an error that appeared in the caption of July's Looking Back section (p. 144).

The invention of the klystron by the Varian brothers, Russel and Sigurd, occurred in 1937 and not in 1953. (See Russel Varian notebook, July 21, 1937 entries, National Archives, Washington, DC.)

Manfred Thumm

Military Spending

▶▶IN YOUR EDITORIAL and in Fred Levien's article in the September 2001 issue of *Microwaves & RF*, you target the Clinton Administration as being single-handedly at fault for the decline of the US military. Professor Levien may have a professional axe to grind, but you as a journalist and editor of a respected publication should exercise more doubt and ask more questions, because this is not true.

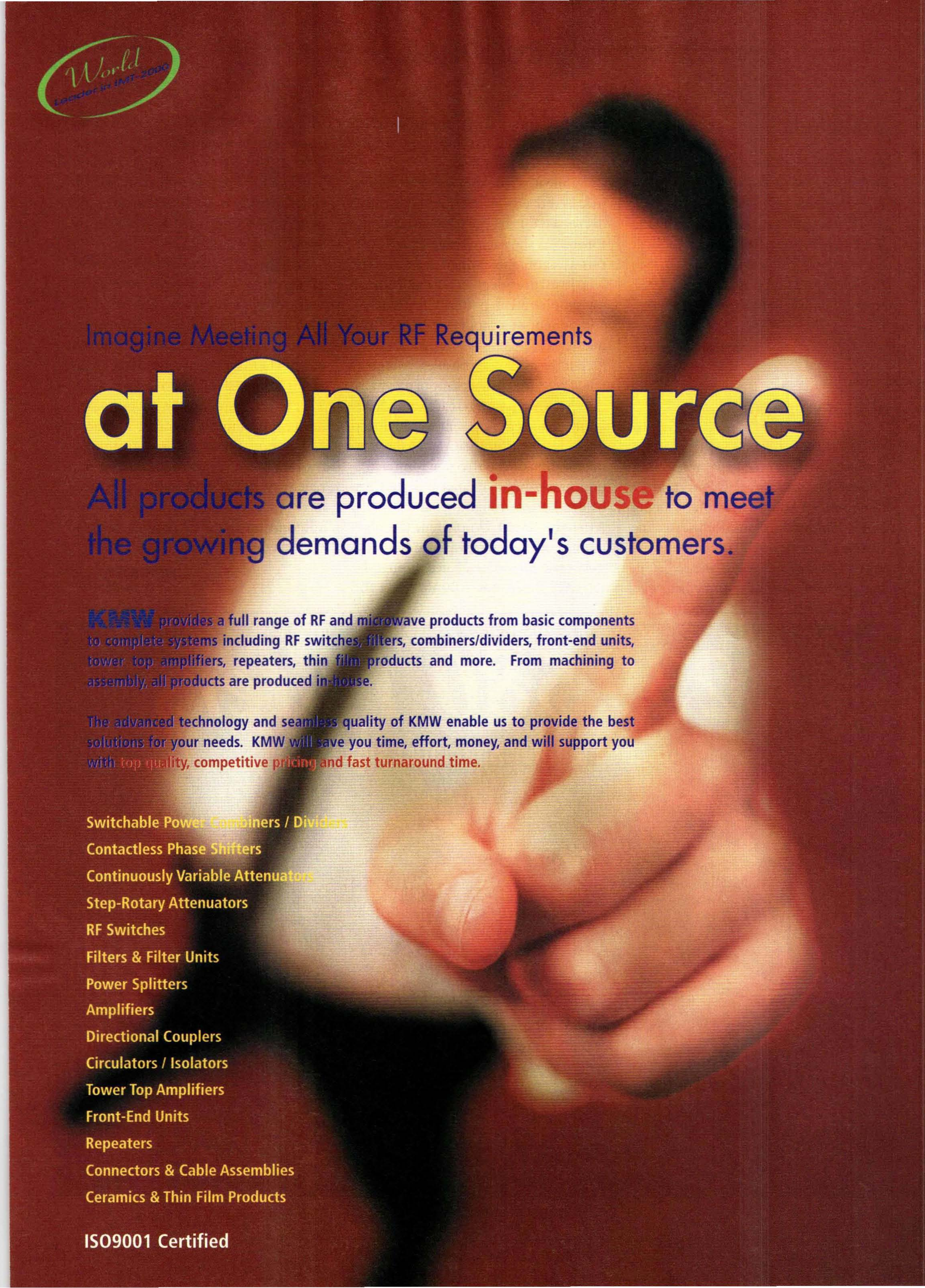
While it is true that social spending increased while military spending decreased during the Clinton administration (as shown in Table 4 on p. 44), this was going to happen regardless of who was president. You can look at the employment patterns in that key component of a modern defense—the microwave business—as proof.

For the most part, business was great until 1988. There were plenty of programs, there was money for research and development, and employment was high and stable. However, in 1988, during the waning months of the Reagan administration, cracks began to appear in the form of program terminations and layoffs. After the Soviet Union fell, the decimation of the microwave and defense industry continued unabated through the G.H.W. Bush years. By the time Bill Clinton arrived in office, entire companies either went out of business or were absorbed by other, larger entities. Much of this consolidation was orchestrated directly from The Pentagon, and would have happened independently of political leadership.

In light of the events after September 11th, while we may be heartened to know that our forces are able to strike back at terrorism and defend our country, it is frightening to realize that much of the industrial base that created the armaments for those forces no longer exists. This is the result of the accumulated neglect and shortsightedness of many.

I do not apologize for any administration, but it seems that many of our recent political and military leaders have failed to heed Thomas Jefferson's admonition that, "The price of freedom is eternal vigilance.

Jesse Sheinwald
Seaford, NY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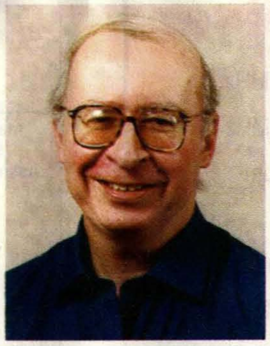
Looking Back And Ahead



REMEMBRANCE AND ANTICIPATION set the tone for many of us during the holiday season. For many, this will be a holiday season like no other, festive but subdued, filled with the memories of September 11th. This has also been a difficult year for many as it has been unstable economically and politically. But when there is hope, there is life. And hope will always feed our anticipation of better things to come, when hatred can be overcome, and when mankind can find peace.

The high-frequency industry survives because of its people, as they learn to work smart rather than hard, and they learn to anticipate future needs.

This year can be thought of as the "interest payment" for the years of prosperity that came before it. In general, companies from IC suppliers to test-equipment manufacturers felt the economic pinch, with many firms forced to trim their workforces through attrition and layoffs. For those companies that rode the wave of prosperity fueled by the growth of cellular and wireless communications during the 1990s, the crash at the end of that ride was particularly painful. The one exception appears to be in software, where suppliers of CAE tools generally enjoyed better-than-average revenue performance during 2001. A possible reason for this is tied to shrinking payrolls: companies that endure layoffs must now ask greater productivity of remaining employees, and good software tools provide a cost-effective means of boosting productivity.

If there is a business lesson to be learned from 2001, it is simply that growth is difficult to manage. During boom years (the 1990s), new orders mean new employees. But during an economic downturn, companies suddenly find themselves with too many people and not enough work. Even companies once considered unstoppable—such as Lucent Technologies and Nortel Networks—can be hurt by the cyclical nature of business.

Every business must endure these cycles and hopefully improve in the process. The microwave industry has shown amazing resilience during its 60 or so years of existence. The industry survives because of its people, as they learn to work smart rather than hard, and they learn to anticipate future needs.

Even during a "down" year, this industry has accomplished a great deal in the way of new products and innovations (see p. 118), and there is much hope for improvement in 2002. The key word is hope and, in keeping it alive, we will survive. From all of us at *Microwaves & RF*, best wishes in the New Year.

Jack Browne

Publisher/Editor

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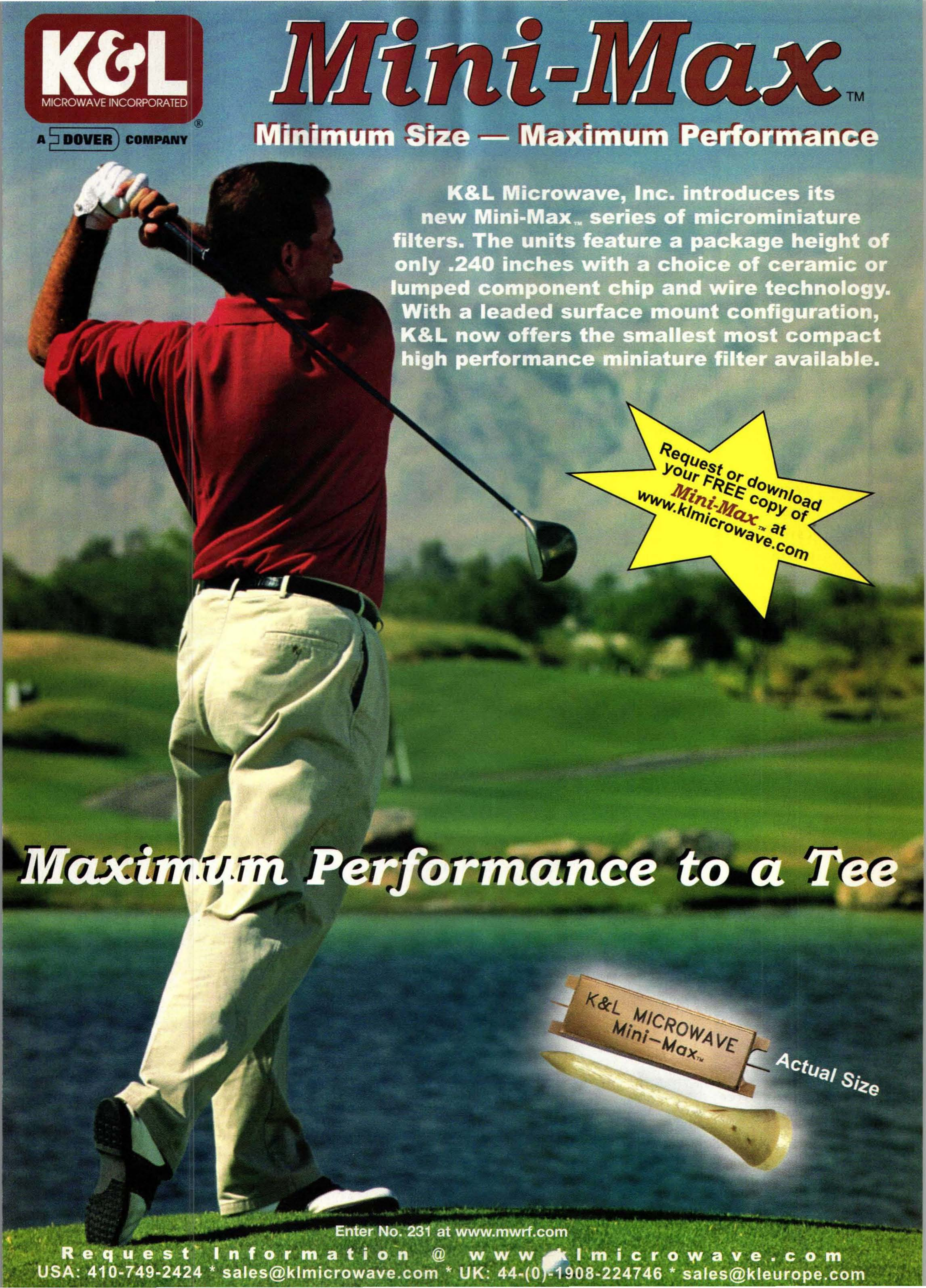
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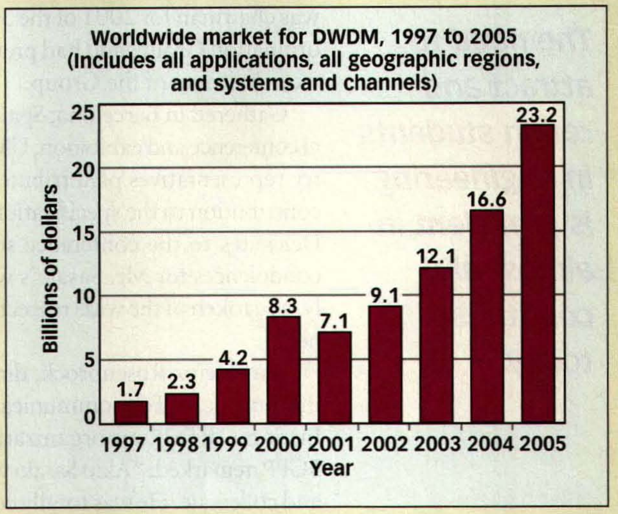
the front end

News items from the communications arena.

Global DWDM Market Will Contract 14 Percent This Year

PROVIDENCE, RI—The worldwide market for dense-wavelength-division-multiplexing (DWDM) transport equipment will decline 14 percent this year to $7.1 billion **(see figure)**, according to *DWDM Markets In Transition*, a report from KMI Corp. The market will resume growth next year, but growth rates will vary sharply by geographic region and by application.



KMI places the strong growth in DWDM spending in 1998 to 2000 and this year's subsequent decline within the perspective of network building cycles. With the Internet boom starting around 1995, bandwidth demand skyrocketed and new long-distance carriers entered the field by building nationwide long-distance networks using the latest optical-fiber and equipment technologies. The market for DWDM systems nearly doubled in 1999 and 2000, from $2.3 billion in 1998 to $4.2 billion in 1999 and to $8.3 billion in 2000.

With most long-distance network buildouts completed by the end of 2000—with capacity sufficient to meet growing demand—greenfield equipment deployments slowed dramatically in 2001. Carriers typically light only a fraction of capacity on newly deployed systems, so equipment installed in one year will not be filled for several years. With the massive deployments that occurred in 1999 and 2000, 2001 and early 2002 will be slack times as carriers simply fill up unused capacity with channel-card deployments.

War On Terrorism To Increase Defense Electronics Budget

ALEXANDRIA, VA—US military spending for electronics and electronic components will increase substantially over the next decade, say experts from the Government Electronics and Information Technology Association (GEIA) in Arlington, VA.

"GEIA forecasts significant new funding for sophisticated combat and combat-support material such as command and control; communications; computers; and intelligence, surveillance, and reconnaissance systems," states a GEIA announcement.

Primary drives in anticipated military and space funding increases include strengthening homeland security, carrying out a war on terrorism, and moving to new 21st century military and aerospace challenges, says the GEIA.

Association leaders have assembled their annual 10-year forecast of markets, programs, and budgets of the US Department of Defense (DoD) and National Aeronautics and Space Administration (NASA). The forecast was presented from October 23 to 25 at the Radisson Hotel in Alexandria, VA.

"We expect that companies with expertise in advanced electronics technologies will be called upon to provide weapons, munitions, and services significantly beyond those planned in the amended 2002 budget," says Dan Heinemeier, president of the GEIA.

"We also project a strong, continuing emphasis on information superiority and information assurance—areas that inherently rely upon considerable electronic content," Heinemeier says.

"The need to attract and retain students in engineering is prevalent in almost all countries today."

Akio Sasaki Is Mourned By The 3G Industry

SOPHIA-ANTIPOLIS, FRANCE—The third-generation (3G) mobile-telecommunications industry has been mourning the death of Akio Sasaki, chairman of the 3G Partnership Project (3GPP) Co-ordination Group. Mr. Sasaki died on October 5, following a stroke that he suffered in July. He was 55 years old.

Akio Sasaki had been managing director of the Japanese Association of Radio Industries and Businesses (ARIB) since October 1999. He was chairman for 2001 of the 3GPP Project Co-ordination Group, and had previously served as vice-chairman of the Group.

Gathered In Barcelona, Spain for their annual conference and exhibition, UMTS2001, industry representatives paid tribute to Mr. Sasaki's contribution to the specification of 3G systems. Delegates to the conference signed a book of condolences for Mr. Sasaki's widow and family, as a token of the wide respect in which he was held.

Karl Heinz Rosenbrock, director general of the European Telecommunications Standards Institute (ETSI), an organizational partner in 3GPP, remarked, "Akio Sasaki was a good friend and colleague. He was totally committed to seeing 3G become a reality, and was a driving force behind the creation of 3GPP in the first instance. We shall miss him greatly."

New Engineering Programs Stress Teamwork And Creativity

NEW YORK, NY—Despite a surge in the overall college-student population and a burgeoning demand for technically trained professionals, engineering enrollments are flat. Since a peak in 1986, the number of engineering bachelor's degrees has declined by 19 percent, according to data from the Engineering Workforce Commission of the American Association of Engineering Societies in Washington, DC. During the 1990s, degrees in electrical and electronic engineering suffered the most serious drop, from approximately 20,000 to less than 13,000. Although certain subfields like computer science and biomedical engineering have seen strong growth, others, such as aerospace and nuclear engineering, now graduate fewer than half as many students as they did 10 years ago.

Things are just as bad in other industrialized countries, including Japan, Germany, and the UK. "The need to attract and retain students in engineering is prevalent in almost all countries today," observes Michael S. Wald, editor of the *International Journal of Engineering Education,* which is based at Ireland's Dublin Institute of Technology. Ireland "is desperately in need of many thousands of engineers and IT [information-technology] specialists," Wald adds.

Reform has been slow to arrive. Legislative and administrative constraints work against it, says Wald. In the European Union (EU), for example, countries have been reluctant to recognize that others also turn out good engineers. Even so, Wald adds, the EU, through a program known as the European credit-transfer system, is working hard "to harmonize engineering education, make sure that degree programs are compatible, and ease the transfer of students between countries and institutions."

Change is occurring in the US as well. Last year, the Accreditation Board for Engineering and Technology (ABET) in Baltimore, MD unveiled its new criteria for evaluating US engineering schools. Rather than simply tallying the courses that students take, the criteria now focus on a student's mastery of specific concepts and processes. The hope is that this approach will free up departments to design courses that are more current and effective.

Microsatellite Is Launched For US Air Force

GUILDFORD, UNITED KINGDOM—PICOSat, a 67 kg microsatellite developed for the US Air Force (USAF) Space Test Program (STP) by Surrey Satellite Technology Ltd. (SSTL) in the UK, was launched successfully from Alaska on September 30.

The PICOSat mission is demonstrating the viability of using a commercial-off-the-shelf (COTS) spacecraft platform to provide cost-effective and timely space flights for Department of Defense (DoD) experiments. This is the first time that the DoD has purchased an "off-the-shelf" microsatellite, which has been tailored by SSTL to carry four experimental payloads for the US Government. It is also the first time that the STP has purchased a spacecraft outside of the US.

The US DoD's objective is to achieve faster mission response and turnaround, cheaper life-cycle mission costs, and more streamlined program execution.

Restoring Phone Service At Ground Zero

MORRISTOWN, NJ—Aiding with setting up portable transmitters/receivers (Txs)/(Rxs) in the World Trade Center disaster zone in lower Manhattan, Edwards and Kelcey, a design and engineering firm, is helping rescue teams communicate and return wireless phone service to normal.

Moving quickly as part of the AT&T Wireless recovery team, the firm brought in several cell sites on wheels (COWs) to the stricken area after permanently installed cell sites were damaged or destroyed. Some permanent cell sites were knocked out immediately following the September 11th terrorist attack.

"The first goal of the team was to position the COWs efficiently to allow local cellular operation recovery at Ground Zero."

"Our recovery team was able to get to the site only hours after the attack to conduct a thorough inspection," says Tom Smith, an executive with Edwards and Kelcey. Smith led the firm's approach team convoy along with Ted Bartlett, regional manager for Edwards and Kelcey's wireless unit.

"The first goal of the team was to position the COWs efficiently to allow local cellular operation recovery at Ground Zero," Smith explains. "We also used portable listening devices to pick up any 911 calls coming from the wreckage," continues Smith.

The team placed mobile cell units at strategic locations throughout the crash site to restore service in the World Trade Center disaster area and in New York's financial district.

COWs are commonly used to provide additional wireless phone capacity in times of natural disaster, such as hurricanes, and at special events such as the 1996 summer Olympic Games in Atlanta, GA.

Kudos

IPC announced that Dieter Bergman, IPC director of technology transfer, has been honored with the Marsh Award at the PCB Design Conference East in Worcester, MA. The Gene Marsh Award for Design Innovation recognizes individuals for significant contributions to the PCB engineering and design industry and is sponsored by *Printed Circuit Design* magazine...ANADIGICS has been awarded a patent for a newly designed GaAs multiband amplifier. The patent, US patent No. 624986 entitled Multiple-band amplifier, is the second one to be granted to ANADIGICS for this particular amplifier circuit. The new GaAs Multiband Amplifier Circuit, which has been designed for use in the wireless communications sector, is disclosed for operation at either the 800- or 1900-MHz band, and provides the desired gain as well as input/output impedance... Sirenza Microdevices announced that it has been registered to ISO 9001-1994 by the Quality Management Institute (QMI) of Canada...LBA Group, Inc. has been granted a patent for its CoLoPole(TM) wireless antenna colocation system. The CoLoPole system permits trouble-free wireless antenna colocation on AM broadcast antenna towers. CoLoPole helps resolve the critical requirement for wireless antenna space by helping to make thousands of AM broadcast towers available for industry use...Galtronics, which is a wireless solutions firm, announced that it has been granted a US patent (No. 6,236,369) for a unique retractable antenna that improves the design of retractable antennas and reduces their cost...RF Micro Devices, Inc. announced that it has again been recognized as one of the fastest-growing technology companies in North Carolina. During the past fiscal year, RFMD posted the sixth-highest five-year growth rate in the state—achieving an increase of 2937 percent. In recognition of that growth, the company has received a "North Carolina Technology Fast 50" award as part of a national program sponsored by Deloitte & Touche, LLP. RFMD has been ranked in the N.C. Fast Tech 50 for the past seven years—receiving first-place honors in 1999 as well as 1997...Labtech announced that HRH The Princess Royal visited Labtech's flagship facility in mid-Wales on October 12th in order to present them with their Queen's Award for Enterprise 2001. The Queen's award for enterprise is the highest honor that can be bestowed on a UK company...International Crystal Manufacturing has met the standards required for ISO 9002 certification and has received their registration number...Palomar Technologies, a manufacturer of automated wire and die-bonding systems for broadband communications, has won the 2001 Advanced Packaging Award for its Laser Diode Attach (LDA) Automated Assembly Cell. The award was sponsored by *Advanced Packaging* magazine. Palomar's entry won the award for the most innovative product that was in the category of die placement and attach, including the handling, alignment, along with attachment of a chip to a substrate...Methode Electronics, Inc. announced that it has been awarded a place in the Deloitte & Touche year 2001 "Technology Fast 50" list for the greater Chicagoland area. MRF

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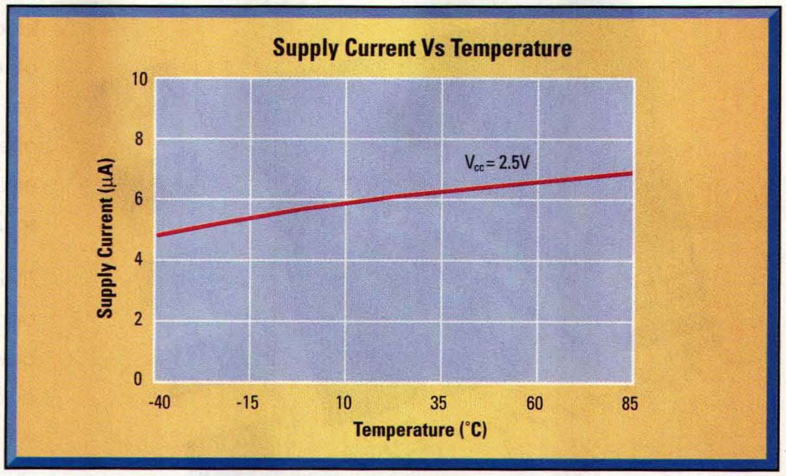

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LM3723	No	N/A		X	X		Push Pull
LM3724	No	N/A		X	X		Open Drain
LM3700/1	Yes	N/A				X	Push Pull
LM3702/3	Yes	N/A		X		X	Push Pull
LM3704/5	Yes	N/A		X	X	X	Push Pull
LM3706/7	Yes	Yes				X	Push Pull
LM3708/9	Yes	Yes		X		X	Push Pull
LM3710/1	Yes	Yes		X	X	X	Push Pull
LM3712/3	Yes	Yes	X	X	X		Push Pull
LP3470	Adj	N/A		(POR)			Push Pull

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Wireless Systems Design Conference & Expo

Tenth Annual Wireless Show Is Renamed And Revamped

Gene Heftman
Senior Editor

Mitchell Gang
Copy Editor

This year's show takes a different road in its approach to keeping designers up to date on technological developments in wireless communications.

Wireless technology keeps moving and changing direction, so the trade shows and conferences that keep design engineers and managers abreast of the latest developments in the industry should as well. That is exactly what is happening to the tenth annual Wireless Symposium, which is changing its name and shifting its emphasis for this year's conference and exhibition. The Symposium's new name is the Wireless Systems Design Conference and Expo 2002, and its workshops and technical sessions offer a more focused as well as system-level perspective view of communications technology than shows of the past. Two things that have not changed are the location—the San Jose Convention Center (San Jose, CA)—and the dates, February 25-28, 2002.

This year's technical tracks are divided into eight categories: Bluetooth/Short Range Communications, Broadband Fixed Wireless, Handset Design, Software, Wireless Internet, Wireless LANs, Wireless Modeling/Test & Measurement, and Wireless Networking. Also on the program are six full-day workshops, some of which are staples from the past (Oscillator Design) and a newcomer, Digital Mobile Radio Fundamentals. Along with picking up some new information on communications design theory, attendees can expect a bustling exhibition hall with many new products on display, some of which are highlighted on the following pages.

As in years past, Randy Rhea of Eagleware Corp. (Norcross, VA) will deliver his Oscillator Workshop on February 25th from 9:00 am to 5:00 pm. This course is aptly called "The Demystification Of Oscillators." Rather than designing by modifying existing designs, the emphasis is on understanding how to design from the ground up. Topics will include phase noise, nonlinear behavior, tuning, quality factor (Q), as well as low and high power. The frequency range of the oscillators spans 100 to 2400 MHz.

As showgoers have heard for the past two years, one day there will be Bluetooth-enabled devices. If this happens to be the year, designers should sit in on the "Bluetooth-RF Basics" workshop given on February 25th from 9:00 am to 5:00 pm. Ken Noblitt, North American technical manager for Cambridge

Silicon Radio (CSR; Richardson, TX) will begin with the basics, and will cover the details of the Bluetooth radio and baseband considerations. This course should benefit anyone planning to design Bluetooth capability into a wireless system.

Mobile telecommunications was, of course, the most important technology of the 1990s and still is in the 21st century. It encompasses cordless telephony; paging; private and professional mobile radio; and its largest component, cellular-radio communications. As digital cellular technology moves toward its third generation (3G), the complexity of systems is increasing with the advent of wideband standards such as wideband code-division multiple access (WCDMA) and cdma2000. To assist those with limited background in cellular and digital wireless communications, and those looking toward the 3G future, the workshop

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"Digital Mobile Radio Fundamentals" will be presented by Rick Fornes, an instructor with Besser Associates (Mountain View, CA), one of the leading technical education companies in the country for working engineers.

Steven Best, president of Cushcraft Corp. (Manchester, NH), returns once again with his popular and informative workshop on one of the most difficult technical areas of wireless communications—antennas. In the "Antennas & Propagation for Wireless Communications" workshop on February 25th from 9:00 am to 5:00 pm, Dr. Best will provide a broad introduction into antenna properties, design considerations, and RF propagation. The workshop tackles antennas from the ground up. It begins with the basic definitions and concludes with an actual antenna design using commercially available antenna-design software.

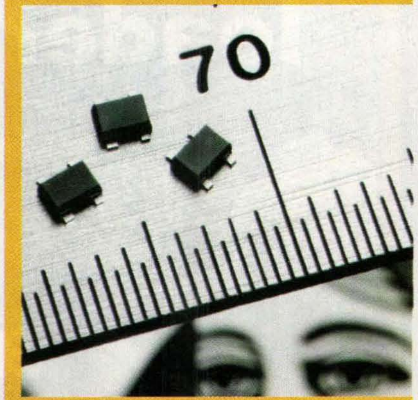
Conference Papers

While almost everyone assumes that Bluetooth will be the short-range RF link for data transfer between portable phones, computers, peripherals, and a myriad of other appliances, it could also be the voice link for the "hands-free" communications concept in cell phones. Bluetooth is not the only choice for the latter. Marc Niklaus, product line manager for XEMICS S.A (Mountain View, CA), will present a paper asking the question, "Does Voice Over RF Transmission Using Low-Cost Devices Operating in the 900 MHz ISM Band Make Sense?" His paper will examine the technical challenges in sending voice over RF and complying with various regional regulations using the 900-MHz industrial-scientific-medical (ISM) band. It will attempt to answer the questions of whether a 900-MHz-based voice solution over RF will be of equal quality to Bluetooth and will it make sense economically.

In the Broadband Fixed Wireless sessions, a paper will address another question about the next generation of wireless systems, "Broadband Wireless—Ready for Prime Time?". The paper suggests that broadband wire-

Transistor Works In 10-GHz Applications

Model NESG2030M04 is a silicon-germanium (SiGe) low-noise transistor that is suitable as a first-stage low-noise amplifier (LNA) or oscillator for 1.6-, 1.9-, 2.4-, 3.5-, or 5.8-GHz designs. With a third-order intercept point (IP3) of +22 dBm, noise is 0.9 dB and associated gain is 16



dB. Fabricated using the 60-GHz fT SiGe UHS2 wafer process, the unit is operational in applications to 10 GHz. The units are housed in a low-profile, flat-lead M04 package that measures 0.6 3 2.0 3 2.1 mm including the leads. The flat-lead design helps reduce lead inductance for optimal electrical performance. **California Eastern Laboratories, 4590 Patrick Henry Dr., P.O. Box 54964-1817, Santa Clara, CA 95054; (408) 988-3500, FAX: (408) 988-0279, Internet: www.cel.com.**
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Tx Targets CDMA Phone Applications

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vices (PCS) band CDMA phone markets. Manufactured with silicon-germanium (SiGe) technology, the Tx is a one-chip solution that integrates all transmit parts from analog baseband input to RF driver output. The unit's integrated high-power drivers support the single-ended and differential output signals. The on-chip auxiliary phase-locked loop (PLL) [including an active voltage-controlled oscillator (VCO)] and the three-wire, 4-b bus programming help to simplify the design in. The Tx is packaged in a 5 × 5-mm lead-frame outline package with 32 pins and evaluation boards. **Atmel Corp., 48 Blvd. Jean-Jaures, 92110 Clichy, France; 33 1 47 30 71 80, FAX: 33 1 47 30 01 89, e-mail: earchambeaud@wanadoo.fr, Internet: www.atmel-wm.com.** Enter **No. 66** at www.mwrf.com

PA Offers 17 dB Of Gain

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less access to the home has not lived up to expectations due to poor performing first-generation (1G) equipment that are comprised of little more than upbanded cable-modem technology. The paper claims that second-generation (2G) equipment brings the promise of non line-of-sight operation, upgradable residential modems that are managed by the service provider and transparent to the user, and low or no installation cost low-wall mounted and/or indoor antennas. The paper will explore how the market will evolve and its impact on the delivery of broadband services to the world.

If 3G technology is to become a technical and commercial success, a number of engineering challenges loom. One problem is providing high data rates to the user in a mobile environment. A paper entitled, "Implementation Considerations of a Multiuser Detection Adaptive Array Receiver for the Uplink of 3G" by Michael LeFevre and Peter Okrah of Motorola's Wireless Infrastructure Division (Arlington Heights, IL) says that sophisticated signal processing is required to overcome the wireless challenge of high data rates in a mobile environment (the target goal is 2 Mb/s). The authors believe that smart antennas and multiuser detection (MUD) are two techniques that can provide the answers. Although the two techniques were developed independent of one another, the authors believe that they can be combined in a system to obtain better signal-to-noise ratio (SNR) and eliminate the phenomenon known as multiple-access interference (MAI). This paper is one of seven offered in the Handset Design technical sessions.

As the world of communications and computers draw closer together, the question arises as to what parts of the computer world fit best with communications applications. A partial answer is offered in the Software sessions by the paper, "Evaluating the Performance of Embedded Java in Wireless Systems" by Markus Levy, president of the Embedded Microprocessor Benchmark Consortium (EEMBC) [El Dorado Hills, CA]. This mini-tutorial will address how to evalu-

ate embedded Java including topics such as Java virtual machines, just-in-time Java compilers, the Java accelerator or coprocessor, and the interaction of these mechanisms with a host operating system in an embedded application. Java may play a role in certain wireless applications, but it will depend on whether the application can run fast enough and how much power it consumes. This session will explain how to develop the proper benchmarks for Java performance and to

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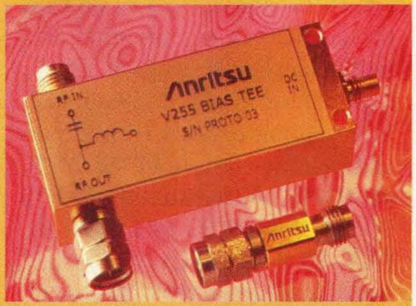

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decide if these meet the requirements of a specific design.

Following the theme of computer/communications convergence is the paper entitled, "Incorporating Wireless Connectivity into Handheld Computers," delivered by Doug Grant, business development manager for Analog Devices, Inc. (Norwood, MA). Presented in the Wireless Internet sessions, the paper suggests that the convergence of handheld computers and the Internet has been held back by the lack of wide-area connectivity. The author's fundamental claim is that it does not make sense to have a portable computer that requires a wired connection to access the Internet. He goes on to compare the alternatives for making wireless connectivity a reality.

Also in the Wireless Internet sessions is a paper by Chung Liu, chief technology officer and VP of engineering for

formance of stainless-steel coupling nuts for applications where many couplings and decouplings are anticipated, with the economy of machined-brass connector bodies. The connectors feature gold (Au)-plated contacts and Teflon insulation. **RF Connectors, 7610 Miramar Rd., San Diego, CA 92126-4202; (800) 233-1728, (858) 549-6340, FAX: (858) 549-6345, e-mail: rfi@rfindustries.com, Internet: www.rfindustries.com.** Enter **No. 69** at www.mwrf.com

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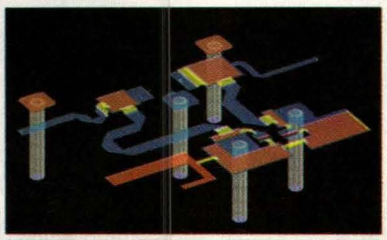

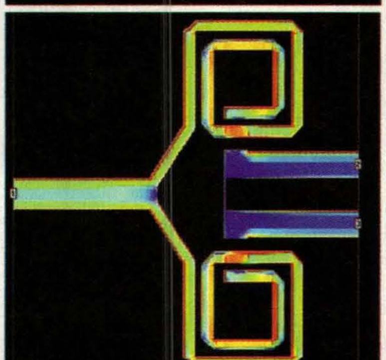

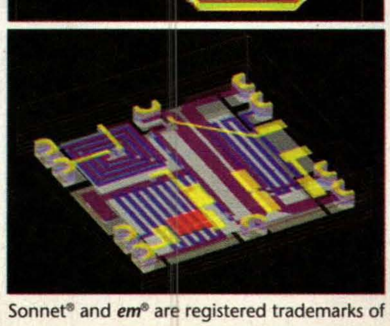

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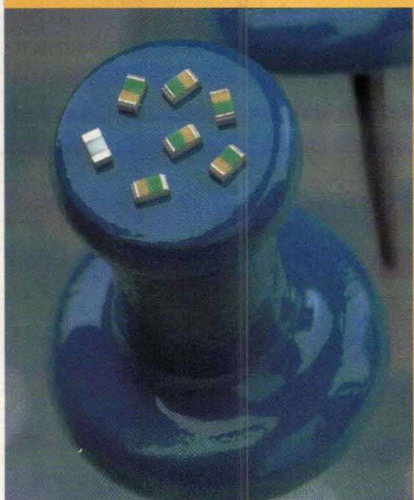
Inductor Spans –55 To +125°C

The Model LLP1005FH is a thick-film chip inductor that was manufactured with a proprietary ceramic material and photolithographic process. Designed for use in high-precision impedance-matching applications for power amplifiers (PAs), the inductor features ±0.2-nH inductance tolerance over the full temperature



range of –55 to +125°C and a temperature coefficient of +100 PPM/°C. Inductance tolerance of ±0.1 nH is available across the inductance range of 1.0 to 2.7 nH. Self-resonant frequency is controlled within ±15 percent, and inductance and quality-factor (Q) values are specified at 800 MHz as well as 100 MHz. Available in 1-to-10-nH inductance values, the inductors are housed in a 1.0 3 0.5 3 0.4-mm package. P&A: 10,000 qty. (tape and reel). **Toko America, Inc., 1250 Feehanville Dr., Mt. Prospect, IL 60056; (847) 297-0070, FAX: (847) 699-1194, e-mail: info@tokoam.com, Internet: www.tokoam.com.**
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ACCESS Systems America, Inc. (Fremont, CA). Entitled "Roadmap from WAP 1.x to 3G and Beyond," it claims that the convergence of Wireless Access Protocol (WAP), i-mode, and mainstream Internet has opened numerous options—architecture, product, vendor—for wireless operators to build on or expand their investments. But Liu warns of pitfalls and dead ends that must be avoided to be successful. Among his suggestions for success are to adopt an open end-to-end

Internet architecture, watch out for the real costs of infrastructure, invest in an infrastructure that enables a return-on-investment, and more

A practical paper opens the Wireless LAN session. Entitled "Troubleshooting Common Wireless LAN Radio Problems," and authored by Richard Abrahams, senior principal engineer at Intersil, Inc. (Mountain Top, PA), it stems from the growing use of wireless local-area-network (WLAN) radio cards in computers. Since many computer companies have limited RF experience, they tend to use reference designs (or reference radios) to speed up their time-to-market capability. A problem arises, however, if a company tries to modify a reference radio to adapt it to a specific application or feature in the end product. Lack of RF expertise leads to problems that computer designers are not equipped to handle. The author discusses the most common WLAN transmitter (Tx) and receiver (Rx) problems and their solutions to enable engineers to troubleshoot much faster. He presents an orderly way to conduct diagnostic routines that will enable engineers to quickly zero in on the problem.

Direct conversion or zero-intermediate-frequency IF (ZIF) radios have been mentioned as a substitute for the conventional superheterodyne types over the last few years, mainly in wireless handsets. A paper by Carl Andren, senior systems engineer at Intersil, Inc. suggests that the architecture is ready to move into WLANs. In "Zero-IF Technology for Low Cost Wireless LANs," Andren says that problems with ZIF can now be solved through integrated-circuit (IC) design and that the technology is ready for mainstream applications. If true, ZIF could simplify WLAN Rxs and make them less costly since the technique eliminates many of the discrete components needed in a conventional superheterodyne Rx.

New Themes

New to this year's show are a pair of tracks that emphasize two of the more complex aspects of wireless technology: antenna design and system testing. One track is known as Wireless Modeling/Test & Measurement, and the other is Wireless Networking.

A paper in the first track entitled "Understanding the Performance Properties and Trade-Offs of Fractal Antenna Designs," by Steven Best, explores the fractal antenna which exhibits the characteristics of resonance compression, multiband behavior, and a lower resonant frequency than conventional Euclidean

Inductors Target Handheld Applications

Model LPO1704 "Power Wafer"Y inductor series is suitable for size-critical applications such as PC cards, personal digital assistants (PDAs), notebook computers, mobile phones, and other handheld devices. System models in the series have inductance values ranging from 1 to 330 µH and root-mean-square (RMS) current ratings up to 3.6 A. The series features a ceramic case that protects the winding, while the case's flat top provides a surface for pick-and-place handling. The inductors are housed in a 6.6 3 5.5 3 1.0-mm package. P&A: $0.49 (10,000 qty.); stock. **Coilcraft, 1102 Silver Lake Rd., Cary, IL 60013; (847) 639-6400, FAX: (847) 639-1469, e-mail: info@coilcraft.com, Internet: www.coilcraft.com.**
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antennas of the same overall size. The size reduction afforded by a fractal antenna is beneficial in antenna design of wireless devices since space constraints are often critical and limit the size of Euclidean antennas. A focal point of the paper is a comparison of the relative performance of fractal and Euclidean antennas that are designed to be resonant at the same frequency. In some cases, the fractal antenna offers advantages over Euclidean types, but in others, the fractal provides little or no performance advantage.

On the Test & Measurement front, Marta Iglesias, marketing engineer at Agilent Technologies (Palo Alto, CA) will present the paper, "Understanding 3G Modulation Quality Measurements." The subject matter pertains to the various ways of characterizing the performance of 3G CDMA base-station Txs. The significance of modulation quality factors such as rho, error-vector magnitude (EVM), code-domain power, peak code-domain error, and the relationships among them are included. Iglesias will also explain a number of additional measurements that, while not part of standard conformance testing, offer engineers more insight into Tx subsystem and system design. The benefits of each measurement will be discussed from the engineering perspective—RF, baseband, and systems integration.

Along the same lines as the previous paper is "Analyzing W-CDMA Signals to Assure 3G Mobile Station and Chipset Performance," by Carla Slater, product marketing engineer at Anritsu Corp. (Richardson, TX). The author points out that WCDMA (which is expected to be the most prevalent 3G technology) will place far greater testing demands on designers than yesterday's 2G technology. The reason, of course, is the fact that 3G will send and receive not only voice, but data and video images with signals that are more complex than 2G. In addition to this is the requirement for WCDMA to coexist with Global System for Mobile Communications (GSM), the world's most widely used wireless technology. For example, 3G systems must be able to verify handovers from WCDMA to GSM

equipment. Another requirement is the ability of WCDMA to communicate with Integrated Services Digital Networks (ISDN). Due to the diversity of these and other applications and their complexity, manufacturers must be able to simulate and conduct a number of challenging tests at the research-and-development (R&D) level.

In the Wireless Networking track, Chris Fisher of XtremeSpectrum (Vienna, VA) will provide attendees with a glimpse into the future of a technology that has received little attention thus far—ultra wideband (UWB). His paper is entitled, "Ultra-Wideband Technology And Its Applicability As A Wireless Networking Technology." UWB operates across a wide frequency spectrum by transmitting a series of extremely narrow (10 to 1000 ps) low power pulses. A UWB Tx can distribute its energy over the equivalent of 1000 TV channels, 30,000 frequency-modulation (FM) channels, or 500,000 walkie-talkie frequencies. The UWB signal at any one frequency is extremely small and serves as a suitable indoor networking technology. Compared with Bluetooth, for example, UWB offers equivalent power consumption and cost, but it runs 100 times faster. This makes it capable of handling multiple data streams of video and audio simultaneously. Sometime in the future, UWB may be the vehicle that brings comprehensive wireless networking to consumers.

With more wireless users taking to the airwaves every year, network operators desire to cram as many users as possible into the available spectrum. To accomplish this, they must have reliable and accurate methods for measuring the traffic flow across the network. According to John Arpee, co-founder and CTO of Scoreboard, Inc. (Herndon, VA), most of today's methods are highly inaccurate. His paper, "Network Traffic and Data Measurement—Tools for Optimizing Wireless Networks,"presents a way for network operators to operate their networks efficiently and plan for future growth through accurate data measurement. The key is to obtain accurate information while shutting down cells during nonpeak hours for a minimum time.

The papers reported in this article represent only a portion of the total that are being offered in the technical conference sessions. Moreover, new papers are being added as this article goes to press. For up-to-date information on what will be presented, go to www.wirelesssystems2002.com and click on "Conference at a Glance."

Device monitors RF radiation levels

THE SMARTS II is a dual-sensor, ultra-wideband monitor covering 2 to 100 MHz. The monitor protects industrial and communications workers near Txs and in other areas with increased risk of exposure to unhealthy levels of RF radiation. The unit is designed for any environment and is powered by a +9- or +12-VDC battery or an external +24-VDC power supply. Under normal conditions, the monitor's alarm LED flashes once every 40 s to indicate a safe environment. A continuous, audible, and visible alarm occurs when the monitor detects RF radiation at or above the field-set threshold level above 50 percent of standard (the maximum safe level of RF energy). SMARTS II offers a field-adjustable alarm-threshold level switch to provide early warning against exposure. Threshold level can be set at 10, 25, 35, or 50 percent of standard. P&A: $3000.

Narda Safety Test Solutions, 435 Moreland Ave., Hauppauge, NY 11788; (631) 231-1700, e-mail: NardaSTS@L-3COM.com, Internet: www.narda-sts.com.

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Probes measure impedance to 3 GHz

A FAMILY OF microwave probes, calibration standards, and probe stations for on-wafer impedance measurements to 3 GHz interfaces with Agilent's E4991A impedance/material analyzer, offering an impedance-measurement solution from 1 MHz to 3 GHz. The ability to perform measurements from milliohm and kilohm ranges makes this solution suitable for wireless applications such as Bluetooth, WLANs, and WCDMA systems. Options include extension cables and a connecting plate, reducing the accuracy degradation caused by improper calibration.

Cascade Microtech, Inc., 2430 NW 206th Ave., Beaverton, OR 97006; (800) 550-3279, (503) 601-1000, FAX: (503) 601-1002, e-mail: sales@cmicro.com, Internet: www.cascade microtech.com.

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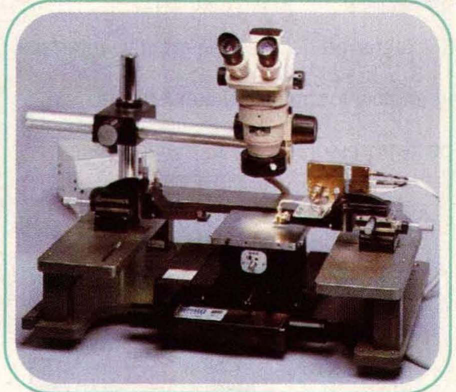

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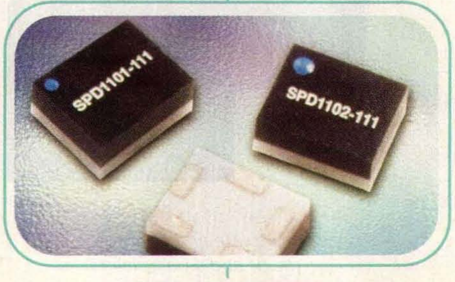

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Detectors target 18-to-22-GHz VSAT applications

MODELS SPD1101-111, SPD1102-111, and SPD1103-111 are surface-mount sampling phase detectors that were designed for use in VSAT and LMDS applications that operate from 18 to 22 GHz. The detectors can be used to phase lock microwave VCOs from 1 to 20 GHz to a stable reference frequency. The detectors were designed with a Schottky series pair, step-recovery diode, and two blocking chip capacitors in a chip-on-board-style package that occupies less real estate than beam-lead versions and enables the use of automated assembly techniques. P&A: $10.00 (1000 qty.);10 wks. ARO.

Alpha Industries, Inc., (978) 247-7700, FAX: (978) 247-7905, e-mail: sales@alphaind.com, Internet: www.alphaind.com.

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Analyzer locates defects at development stage

MODEL D3371 IS a transmission analyzer that operates at data rates up to 3.6 Gb/s and is primarily designed for use in R&D, as well as in the production of components and systems for SDH, SONET, ATM, Gigabit Ethernet, and Fibre Channel. The unit can detect defects and error sources at the development stage so that problems are avoided from the outset. The complementary data output of the D3371 can be set flexibly and has an amplitude resolution of 10 mV and 1 ps in the phase range. The output amplitude range can optionally be extended to +3 VDC. The analyzer provides all of the standard measurement functions of a BERT and offers different evaluation facilities for the variety of available test patterns. Test signals for specific frame patterns are available and can be analyzed as required. Overhead for FEC applications is provided.

Rohde & Schwarz GmbH & Co. KG, D-81671 Munchen, Muhldorfstr. 15; +4989/4129-11765, FAX: +4989/4129-13208, Internet: www.rohde-schwarz.com.

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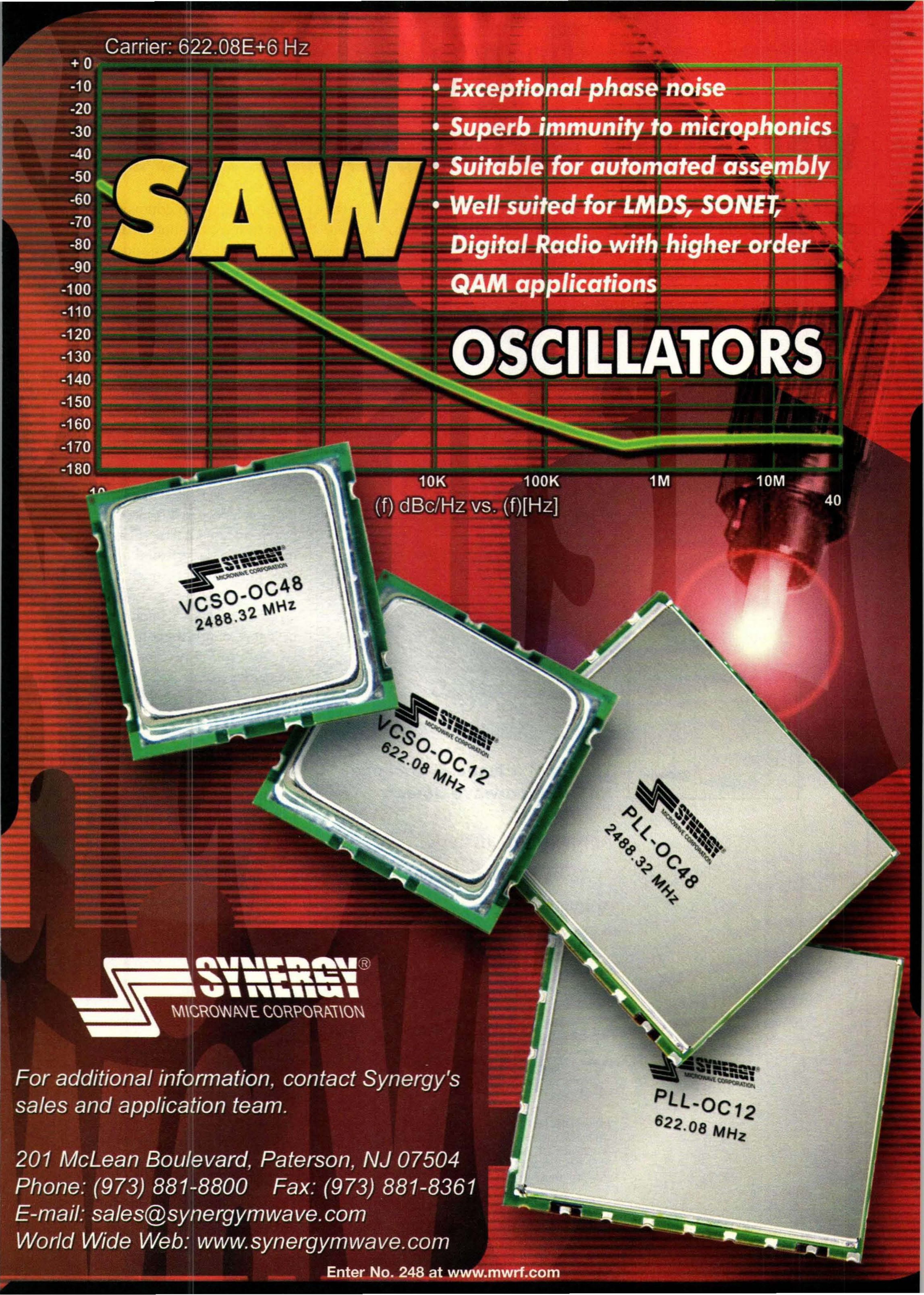
Carrier: 622.08E+6 Hz
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War Spurs Hiring About-Face

In the midst of the sadness and carnage of September 11th, the war on terrorism is having an unintended, yet beneficial consequence on employment in the slumping high-technology sector. According to a *Wall Street Journal* article, while leading high-tech companies in communications, computers, and electronics are laying off thousands of employees, defense contractors are recruiting engineers and technicians to support hardware and software Pentagon projects that are needed for the war effort.

The hiring increases at contractors such as Raytheon, TRW, and General Dynamics will not return defense employment to 1970-1980 levels, but could make a dent in the dismal electronics industry job picture. Some industry insiders have predicted that this year will be the worst ever for employment in the high-tech sector. Projections by outplacement firm Challenger Gray & Christmas put the layoff total at more than 426,000 workers through August. This amounts to more than one-third of the layoffs in the US. Unemployment in Santa Clara County, the core of Silicon Valley, rose to 5.9 percent in September, from 1.8 percent a year earlier, according to the *Wall Street Journal* article.

Uncertainties about the course of the war have set off a global recession that will affect business and hiring in the coming year. High technology is not immune to this situation, but a bright spot could be the demand for more information-technology products by the military. A case in point is the use of Palm, Inc. (Santa Clara, CA) handheld computers aboard naval vessels operating in the sea off of Afghanistan. Sailors use the devices to download e-mails, access the ship's plan of the day, and perform ship-related tasks. Many of the crew brought their own computers with them, but the military has plans to make this technology available to a wide spectrum of military personnel. The Army says it wants to provide its soldiers with "information dominance." If this idea permeates the entire military, the hardware and software required could generate enough business to perk up a high-technology industry facing a grim short-term outlook. MRF

CONTRACTS

Mericom Corp.—Has been chosen by Verizon Wireless to expand and enhance its service in Southern Oregon. The $2 million project, to be completed using Mericom's in-house resources, will include approximately 60 sites in the Roseburg, Medford, Klamath Falls, and Bend areas.

Allgon—Has signed a $13 million general agreement with a global mobile operator. According to the agreement, Allgon will provide base-station antennas for use in GSM networks in North America.

Eurolink Ltd.—Has been awarded a contract to develop with the National Association of Radio and Telecommunication Engineers (NARTE) a new certification program in Product Safety.

CTA Communications, Inc.—Was awarded a contract by the US Department of Justice to provide engineering support for a land mobile radio network for the federal agency. CTA will develop a nationwide Justice Wireless Network (JWN) High Level System Design to identify industry standards and other design elements common to all JWN zones across the US. The contract amount is for one year, with four one-year renewal options, with a total contract value of more than $4.6 million.

Andrew Corp.—Signed a two-year antenna manufacturing agreement with Ericsson Microwave Systems AB. Andrew will be one of the manufacturers and suppliers of integrated antennas—as well as Andrew ValuLine® antennas—for use in Ericsson's MINI-LINK™ point-to-point microwave solutions.

Sprint— Announced expanded coverage of the Sprint PCS digital wireless service in Cortland, NY. The $3.25 million project brings new service to population centers in central New York, including Cortland, Homer, Cortland West, much of Munson's Corners, Marathon, Little York, Baltimore, LaFayette, and Prebele, as well as to the highways connecting them. Local Sprint PCS service is operated and managed by Independent Wireless One, a Sprint PCS Network Partner.

Motorola—Announced the signing of a contract with China Mobile Communications Corp. (China Mobile) to expand China Mobile's GPRS network. Motorola was chosen as one of the suppliers to China Mobile's GPRS network last year. Financial details of the contract were not disclosed. The contract calls for the implementation of the Motorola Global Telecom Solution Sector's (GTSS)/Cisco Systems, Inc. GPRS network solution in seven major Chinese provinces and municipalities, including Beijing, Tianjin, Zhejiang, Sichuan, Hunan, Jiangxi, and Liaoning. The agreement also provides that Motorola will significantly upgrade China Mobile's existing GSM networks in the previously mentioned seven regions. Also, 10 other provinces and special municipalities in the country, through the provision of Packet Control Units (PCUs), will be upgraded including system equipment and software upgrades. The expansion program will increase the network capacity in the seven provinces and municipalities by 350,000 subscribers.

FRESH STARTS

Anritsu Co.—Has partnered with Wireless Valley Communications where Anritsu's MS2711A Handheld Spectrum Analyzer will feature Wireless Valley's SitePlanner® engineering systems. The agreement combines a handheld measurement instrument with software packages for designing, measuring, and maintaining in-building, campus-wide, and microcell wireless communications systems.

Ansoft Corp.—Announced an agreement with test-and-measurement equipment manufacturer Rohde & Schwarz to provide links between its popular WinIQSIM communication-waveform-generation software and Ansoft's communication design products—Serenade® and Symphony. The new capability will allow RF designers and system architects to simulate communication systems under the same conditions used in hardware testing and product development.

Independent Wireless One, a Sprint PCS Network Partner—Has completed its switching center in Londonderry, NH, expanding Sprint PCS coverage in the state by more than 50 percent. The $7.5 million investment in the switching facility has created 12 new high-tech jobs in Londonderry, including positions for highly skilled RF engineers and switch technicians. It is part of the ongoing program to expand capacity of the Sprint PCS nationwide network, which involved more than $1.06 billion capital expenditures for the second quarter of 2001 alone.

Cadence Design Systems, Inc.—Has launched www.spectraquest.com, an online community for PCB engineers and designers to learn more about—and collaborate on—high-speed design issues. Topics such as constraint development, simulation, modeling, power delivery-system design, constraint-driven placement and routing, and achieving signal integrity are covered.

Anaren Microwave, Inc.—Has acquired all outstanding capital stock of Amitron, Inc., a privately held North Andover, MA-based company. Amitron is a manufacturer of precision thick-film ceramic components and circuits for the medical, telecommunications, and defense electronics markets.

Accelerated Technology, Inc. (ATI)—Announced the selection of Nucleus software in the PAVIC portable multimedia device by Varo Vision of Korea. The PAVIC provides MP3 music file playing, digital voice recording, and mobile Internet-access functions. It is also equipped with a built-in digital camera. To develop and design the multiple functions of the device, engineers at Varo Vision chose ATI's Nucleus PLUS real-time kernel, Nucleus NET TCP/IP networking protocol stack, Nucleus PPP point-to-point protocol, and Nucleus FILE file-management system.

Pfizer, IBM, and Microsoft—Have launched Amicore, an independent software and services company providing workflow and connectivity solutions to office-based physicians. Amicore's focus will be to reduce the administrative workload and paperwork for physicians. MRF

EMS Names Cook To VP Position

COOK

EMS Technologies, Inc. has appointed LINDA S. COOK as vice president and director of Advanced Extremely High Frequency (AEHF) Projects, within EMS' Space & Technology Group/Atlanta. Cook was formerly director of engineering at EMS Wireless.

Advance Fiber Optics—GERALD KELLY to manager of the OSP InSight software-development group; formerly developed medical-claims-analysis software for Ingenex.

CTS Corp.—JAMES K.C. CHEN to vice president of Taiwan business development; formerly managing director for CTS Components Taiwan.

Manufacturing Technology, Inc. (MTI)—TIM PYLE to West Coast and Southwest manufacturer's representative; formerly worked on MTI's sales efforts.

LBA Technology, Inc.—MARCIAN L. BOUCHARD to president; formerly served as vice president and general manager of Firetrol, Inc., a division of Emerson Electric.

Virginia Tech's Center for Wireless Communications (CWT)—GEORGE E. MORGAN to director; formerly director of the Space and Wireless Business Center.

Point .360—HAIG S. BAGERDJIAN to chairman of the board; remains as executive vice president of Syncor International Corp. and as president and CEO of Syncor Overseas Ltd.

MCE Technologies, Inc.—MICHAEL HUGGAN to managing director of MCE/DML Microwave Ltd.; formerly director of operations at BSC Filters Ltd.

Giga-tronics, Inc.—JOHN R. REGAZZI to vice president of engineering for the Instruments Division; formerly R&D project manager for Hewlett-Packard/Agilent.

Valence Semiconductor—DR. GLENN GULAK to vice president of engineering and chief technology officer; formerly senior technology advisor.

M2 Global Technology Ltd.—ANTHONY L. EDRIDGE to manager of engineering; formerly engineering manager at Data Race, Inc.

Three-D OLED, LLC—TOM MILLER to vice president and general manager; formerly vice president of sales and marketing at Silicon Motion, Inc.

The American Industrial Hygiene Association (AIHA)—RICHARD A. "DICK" STRANO, CAE, to executive director; formerly executive director of the American Conference of Governmental Industrial Hygienists (ACGIH).

Cooper Electronic Technologies—ED CARTER to vice president of sales; formerly served as director of sales for North America.

HARTING, Inc. of North America—RICHARD A. MACK to general manager of the Electronic Business Unit; formerly director of Marketing for the Electronic Business Unit.

MACK

WHIPPLE

Andrew Corp.—DENNIS L. WHIPPLE to the board of directors; formerly chairman and CEO of Evercom Communications, Inc.

Optical Cable Corp.—NEIL WILKIN to CFO and the board of directors; formerly served as CFO and treasurer at Homebytes.com, Inc. MRF

SHORT COURSES

Fundamentals of Wireless Data Communications
January 23-25, 2002 (Madison, WI)
The University of Wisconsin-Madison
Dept. of Engineering Professional Development
Madison, WI 53706
(800) 462-0876, FAX: (608) 263-3160
e-mail: custserv@epd.engr.wisc.edu
Internet: http://epdweb.engr.wisc.edu

Mobile Communications and Wireless Networks
February 12-15, 2002 (Reston, VA)
Learning Tree International
Reston, VA 20190-5630
(800) 843-8733, FAX: (800) 709-6405
Internet: www.learningtree.com

Satellite RF Communications and Onboard Processing
March 5-7, 2002 (Cleveland, OH)
Applied Technology Institute
Clarksville, MD 21029
(888) 501-2100, FAX: (410) 531-1013
e-mail: info@ATIcourses.com
Internet: www.ATIcourses.com

MEETINGS

2002 International CES (Consumer Electronics Show)
January 7-11, 2002
Las Vegas, NV
Internet: www.CESweb.org

Wireless Communications Association 8th Annual Technical Symposium
January 14-16, 2002 (Fairmount Hotel, San Jose, CA)
Wireless Communications Association International
Washington, DC 20036
(202) 452-7823, FAX: (202) 452-0041
Internet: www.wcai.com

SUPERnet 2002
January 21-24, 2002 (Santa Clara Convention Center, Santa Clara, CA)
Telecommunications Industry Association
(312) 559-4600
Internet: www.supernet2002.com

2002 Measurement Science Conference
January 24-25, 2002 (Disneyland Hotel, Anaheim, CA)
Measurement Science Conference
(888) 690-8880
e-mail: johnbowman@fluke.com
Internet: www.msc-conf.com

International Solid State Circuits Society (ISSCC 2002)
February 4-6, 2002 (San Francisco Marriott Hotel, San Francisco, CA)
The Solid State Circuits Society of the IEEE
Piscataway, NJ 08855-1331
Internet: www.isscc.org

6th International Commercialization of Military and Space Electronics & Exhibition
February 11-14, 2002 (Los Angeles Airport Marriott Hotel, Los Angeles, CA)
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Huntsville, AL 35801
Dale L. Stamps
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February 20-22, 2002 (New Delhi, India)
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e-mail: llochow@sshowmgmtservices.com

ICAPS 2002 International Conference on Advanced Packaging and Systems
March 10-13, 2002 (Reno Hilton Hotel, Reno, NV)
International Microelectronics and Packaging Society (IMAPS)
Tim Lenihan
e-mail: TimLenihan@imaps.org
Internet: www.imaps.org.

RF & Hyper Europe 2002
March 26-28, 2002 (Paris Expo Exhibition Center, Porte de Versailles, France)
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2002 IEEE AP-S International Symposium on Antennas and Propagation and USNC/URSI National Radio Science Meeting
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IEEE Antennas and Propagation Society (AP-S) and the International Union of Radio Science (URSI)
Krys Michalski
2002 TPC Chair
Dept. of Electrical Engineering
Texas A&M University
College Station, TX 77843-3128
(979) 845-5203, FAX: (979) 845-6259
e-mail: krys@ieee.org
Deadline for web submissions: January 15

6th International Conference on Signal Processing (ICSP'02)
August 26-30, 2002 (Beijing, China)
The Chinese Institute of Electronics
Prof. Yuan Baozong
Northern Jiaotong University
Beijing 100044, China
Internet: http://icsp02.njtu.edu.cn
Deadline for paper submissions: January 10

Serial-Link Transceiver Boasts 8-GSamples/s DAC/ADC

HIGH-SPEED SERIAL-LINK TRANSCEIVERS are important parts of wide-bandwidth digital communications links, typically used to provide the digital filtering and equalization necessary for maintaining high data rates. Chih-Kong Ken Yang and associates from the University of California at Los Angeles (UCLA) and Stanford University (Stanford, CA) detail a serial-link transceiver that uses a 4-b Flash ADC for the Rx portion and an 8-b current-steering DAC for the Tx. The 8-GSamples/s converters are eight-way time-interleaved units. Digital compensation is used to reduce the input offset of the ADC to less than 0.6 LSB. The eight 4-b Flash ADCs are each synchronized by a phase-shifted 1-GHz clock. Digitally controllable offset adjustments in each of the comparators compensate for offsets due to device mismatch, reference ladder mismatch, and systematic DC noise. The Tx consists of eight time-interleaved 8-b DACs, which are also clocked at 1 GHz. Programmable memory stores the sequence to be transmitted, providing flexibility to explore a wide range of modulation options, equalization, and calibration techniques. The multiple clock phases used for the Rx and Tx are generated from two PLLs which are locked to an external divide-by-four reference clock that is running at a nominal rate of 250 MHz. The PLLs are supported by a low-jitter VCO that is comprised of a ring of four differential buffer stages with the eight internal clock phases tapped and driven to each of the interleaved converters. At a VCO frequency of 1 GHz, which corresponds to 8 GSamples/s, the nominal phase/timing resolution is 8.3 ps. By achieving high timing resolution, the transceiver can support fast data transfers at low BERs. For more information, see "A Serial-Link Transceiver Based on 8-GSamples/s A/D and D/A Converters in 0.25-μm CMOS," *IEEE Journal of Solid-State Circuits,* November 2001, Vol. 36, No. 11, pp. 1684-1692.

Planar Rectennas Aid Wireless Power Transfer

MORE THAN 20 YEARS AGO, legendary scientist and engineer Bill Brown pursued a dream of wireless power transfer based on space-based solar arrays and satellite transmissions while working at Raytheon Co. (Lexington, MA). Today, Jouko Heikkinen and Markku Kivikoski of the Institute of Electronics at the Tampere University of Technology (Tampere, Finland) continue that research through the development of planar rectennas, which are combinations of antennas and rectifiers. The use of a rectifying antenna makes it possible to receive RF power from a remote source and convert it to DC power. The Finnish researchers considered a number of ISM wireless bands from 869 to 2450 MHz for the power transfer, using the Friis power-transmission equation as the basis for their work. Since smaller antennas could be used at the upper-frequency edge of the band of interest, the 2.45-GHz band was selected. Rectennas were then designed with three different PCB materials: FR4 (dielectric constant of 4.3), RT5870 (dielectric constant of 2.35), and RO3010 (dielectric constant of 10.2) laminates. RF/DC conversion circuits were designed using discrete Schottky barrier diodes. Although simulations matched closely with measured results, performance levels (+1-VDC output over a distance of 1 m for +26-dBm transmitted RF power with the RT5870 material) were not yet adequate for commercial systems. For more information, see "Performance and Efficiency of Planar Rectennas For Short-Range Wireless Power Transfer at 2.45 GHz," *Microwave and Optical Technology Letters,* October 20, 2001, Vol. 31, No. 2, pp. 86-91.

Phased-Array Antenna Has Piezoelectric Transducer

PHASED-ARRAY ANTENNAS PLAY KEY ROLES in defense electronics and satellite communications systems. But broadband phased-array antenna systems are typically expensive, limiting their application in commercial communications systems. Fortunately, Tae-Yeoul Yun and Kai Chang of the Dept. of Electrical Engineering, Texas A&M University (College Stations, TX) developed a low-cost design for a phased-array antenna that is capable of operating from 8 to 26.5 GHz. It incorporates a multiline configuration with progressive phase shifts. The array uses a novel phase shifter that is controlled by a PET. A dielectric perturber attached to the PET moves vertically on microstrip lines with DC bias voltage, creating the shift in phase and helping to dramatically reduce the number of phase shifters that are required in the array. For more information, see "A Low-Cost 8 to 26.5 GHz Phased Array Antenna Using A Piezoelectric Transducer Controlled Phase Shifter," *IEEE Transactions on Antennas And Propagation,* September 2001, Vol. 49, No. 9, pp. 1290-1298.

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DESIGN

Weigh Amplifier Dynamic-Range Requirements

Although signals for digital communications systems may seem simple, amplifier requirements for these signals can be quite complex and demanding of high linearity.

Amplifiers are often characterized in terms of gain, noise figure, and maximum output power when used in analog applications. For modern data-communications systems, however, designers are often more concerned with nonlinearity and distortion levels. Quantities such as input or output third-order intercept points, spurious-free dynamic range, composite second-order (CSO) distortion, composite- triple-beat (CTB) distortion, and cross-modulation become critical specifications for digital communications systems.

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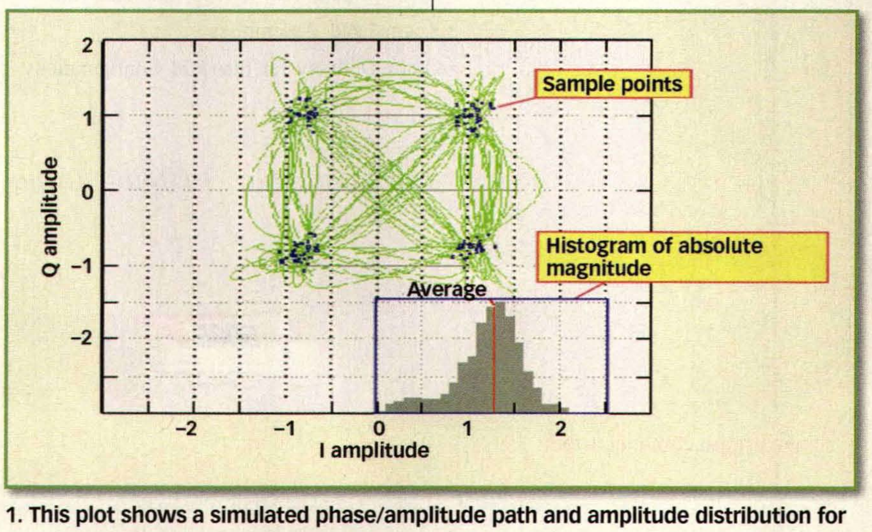
The purpose of digital transmission is to move a sequence of digital ones and zeros from one location to the next, and the representation of intermediate states (the linearity of the system) would seem irrelevant. But linearity is critical to the success of a digital communications system, chiefly due to the limited frequency spectrum available to signals in wireless communications systems, and the complex characteristics of the communications channel that uses that spectrum. Any distortion is equivalent to changes in the Fourier transform of the signal. That is, unintended frequencies are radiated which may interfere with neighboring channels. Thus, radio designers must avoid



1. This plot shows a simulated phase/amplitude path and amplitude distribution for a random digital bitstream with QPSK modulation.

operating at amplitudes at which such distortion is significant.

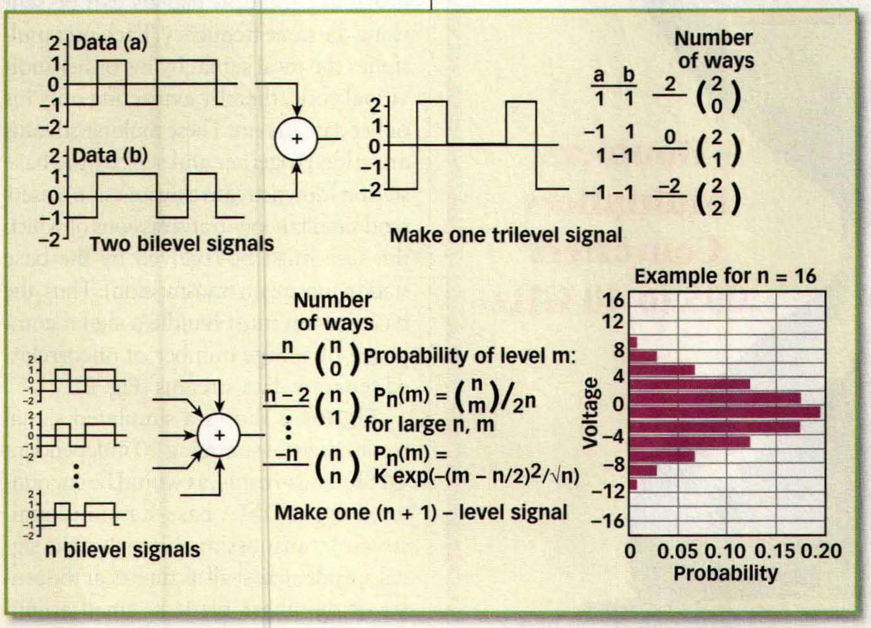
Due to the typically large ratio of the peak-to-average power levels of digital signals (the crest factor), low-distortion transmission of digital signals can become quite complex. A large crest factor can lead not only to interference with adjacent channels, but to additional in-band distortion, and consequent increases in the bit-error rate (BER).

To make efficient use of available spectrum, wireless communications channels filter digital data to smooth the transitions between bits as much as possible while maintaining the integrity of the signal at the "sample points."[1] Modulation schemes, which transmit more than one bit per symbol, are used to maximize the data rate for a slice of spectrum at a given signal-to-noise ratio (SNR).[2] The net result is that the actual analog signal transmitted is considerably more complex than the binary sequence it represents. **Figure 1** offers an example, a quadrature-phase-shift-keying (QPSK) signal. Shown is the modeled path of a filtered QPSK signal, composed of two pseudorandom bitstreams, represented in the phase/amplitude plane. Sample points (i.e., the location of the signal in the I/Q plane at the sample times) are shown, and a histogram showing relative probability of various values of signal amplitude.

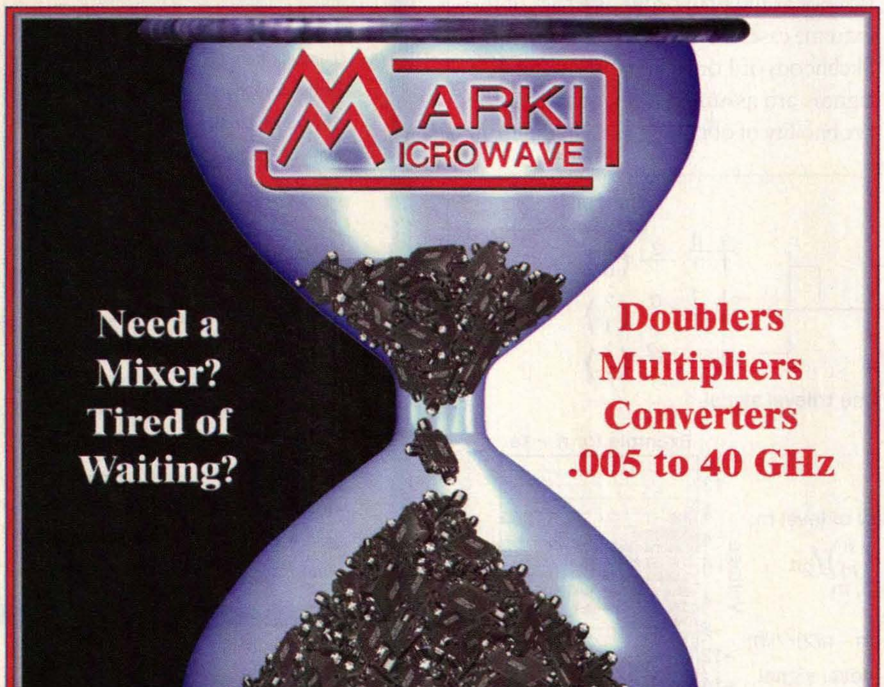
The histogram shows that the average value of the signal amplitude is approximately 1.3 (in normalized units), whereas rare excursions to much larger amplitudes also occur. The ratio of the peak power to the average power for the QPSK signal is approximately 4.3 dB (a factor of 2.7) where the peak is taken from a sample of 256 trajectories (sample to sample) and, thus, is at a probability level of roughly 10^{-3}.

In addition to the challenges of filtered complex constellation paths, real signals often transmit more than one channel simultaneously. This inevitably leads to peak levels that are sometimes much higher than the average signal power. **Figure 2** shows this schematically: the addition of multiple uncorrelated bitstreams produces a final signal which can have many possible levels.

The probability of each "voltage" level is proportional to the number of ways in which that voltage can result. For example, in the case where two bilevel signals are combined, there is only one way to make a level of +2 and only one way to make a level of –2, but there are two ways to make a level of 0. Thus, a zero is twice as likely to occur as either of the extreme cases. In the general case, if equal likelihoods of 1 or –1 in the incoming data signals are assumed (see equation), the probability of obtaining a level "m" from



2. The sum of a large number of binary signals gives rise to a normal distribution.



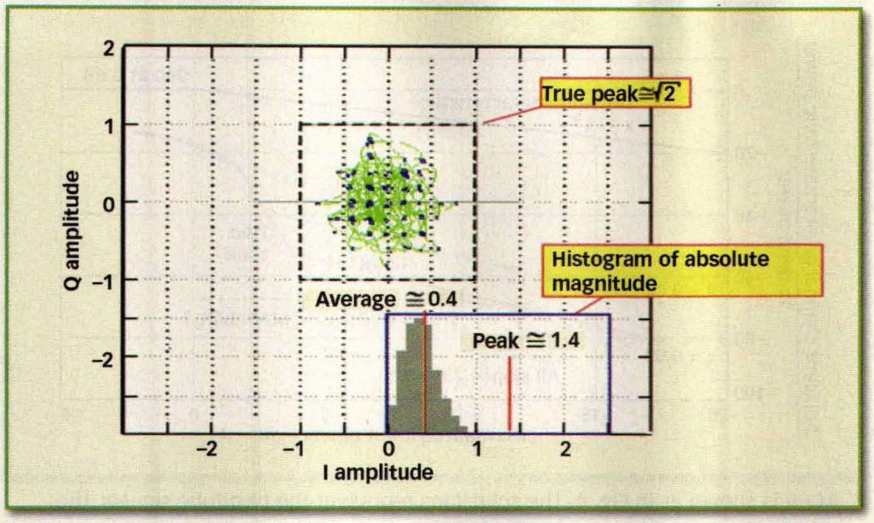
n bilevel signals is described by the binomial coefficient:

$$\binom{n}{m} = \frac{n!}{m!\,(n-m)!}$$

As the number (n) of signals grows large, the distribution of voltages closely approximates a normal or "Gaussian" distribution, with standard deviation of N/2. The distribution of signal power is a chi-squared distribution of order 1 for the case where each signal is either on or off, or of order 2 in the case where there are two orthogonal components, the in-phase (I) and quadrature (Q) signals, combining to form the final signal. The peak-to-average power ratios of the order-1 and order-2 signals are 12.8 and 13.3 dB, respectively, at a probability of 10^{-5}.

A practical example of the superposition of multiple uncorrelated data streams to produce a complex analog signal is encountered in code-division-multiple-access (CDMA) schemes used for mobile communications. In this approach, signals can be sent to many users at the same time using the same frequency without significant interference. Each user's binary bit stream is multiplied by a code consisting of a sequence of very short "chips." The resulting signal has a higher effective bit rate and is thus spread out in frequency, but if codes are chosen to be orthogonal to each other or nearly so, multiple signals can be sent using the same frequency. Each user multiplies the total signal by his or her individual code, thereby extracting only his or her data stream. These multiple streams are added together and sent by the base station (downstream transmission); users send simultaneous transmissions of which the sum must be received by the base station (upstream transmission). Thus, the base station must handle a signal composed of a large number of uncorrelated separate data streams (Fig. 2).

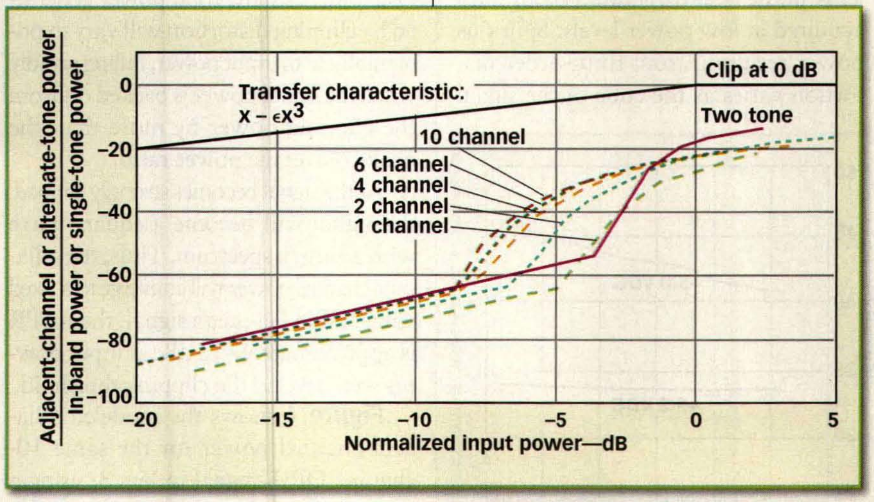
Figure 3 shows a simulated signal resulting from summing 10 independent QPSK datastreams, as would be encountered in a CDMA base-station (downstream) transmission. Note that the signal spends almost all its time near the center of the phase plane at small amplitudes, but at rare intervals an extreme value



3. This plot shows a simulated phase/amplitude path and amplitude distribution for 10 superimposed QPSK-modulated signal streams. Note that in this limited sample of 250 data points, the outer points of the possible constellation are never accessed.

occurs. In this simulation, the peak-to-average ratio is about 6.8 dB (a factor of 5), again for a sample of 256 trajectories, so there is a probability of approximately 10^{-3}. Note that this value is a lower bound on the peak-to-average ratio that would be obtained with a larger sample of points. In particular, the trajectories in this sample never access the corners of the square grid of possible constellation points. (A real signal would go a bit farther, as noted before, due to filtering.) For a more realistic symbol error requirement of approximately 10^{-5}, the peak-to-average ratio would approximately be 11 dB, already close to the limiting values corresponding to a Gaussian distribution in each axis.

Orthogonal-frequency-division-multiplexing (OFDM) modulation as used in fixed wireless communications systems and in proposed wireless local-area-network (WLAN) standards also combines a number of uncorrelated data streams, but at distinct carrier frequencies. The resulting signal displays rare excursions to

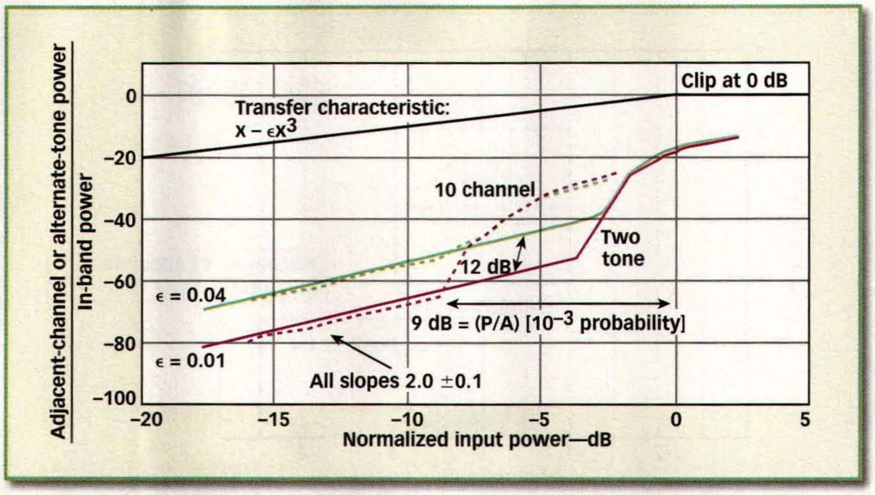


4. Adjacent-channel power ratio (ACPR) is shown as a function of relative input power, for signals constructed from a varying number of QPSK input channels (the dashed lines), and a piecewise-cubic transfer characteristic. The two-tone third-order intermodulation product relative to the tone amplitude (the solid line) is included for comparison.

amplitudes that are much larger than the average and, thus, a large peak-to-average power ratio. Cable-television (CATV) systems combine simultaneous signals from up to 110 separate carriers, often mixing digital and conventional NTSC analog signals. CATV signals are also characterized by peak-to-average power ratios of approximately 11 to 13 dB. (CATV systems also span multiple octaves, so that many distortion products are inband, and represent a significant linearity challenge for the designer.)

The ratio of the power in the adjacent channel to that in the intended channel—the adjacent-channel power ratio (ACPR)—is often a key specification that wireless communications systems must meet to avoid interference between users in different channels. Since distortion is highly dependent on signal amplitude, the ACPR that is expected with a given average signal power depends strongly on the type of signal being sent.

It is important to note that while the ratio of in-band distortion to input signal goes as the square of the input power, the ratio of adjacent-channel distortion to adjacent-channel power goes as the cube of the input power. The in-band channel power increases when the distortion increases, but the adjacent channel is uncorrelated with the channel generating the distortion and, thus, might operate at low power when the interference becomes high, degrading the SNR and increasing the likelihood of bit errors. Thus, it is often the case that even though the inband distortion is larger than the adjacent-channel distortion, it is the ACPR that causes problems with meeting performance specifications.



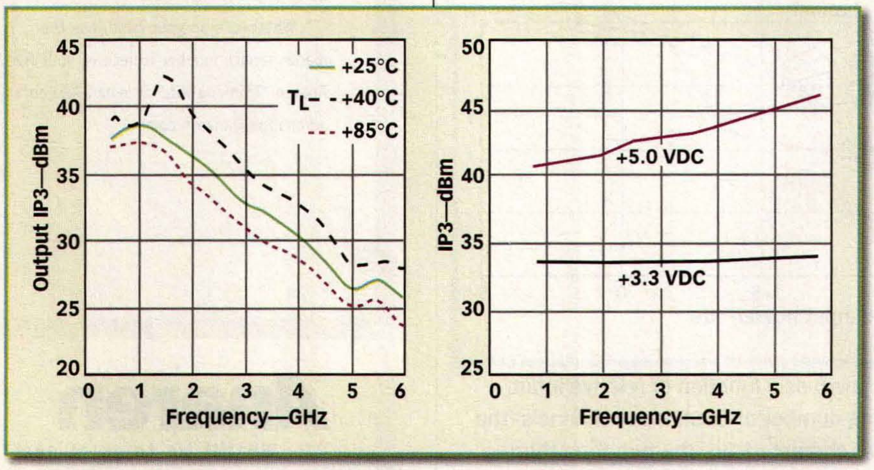
5. ACPR is shown as in Fig. 4. The solid lines represent the two-tone signals, the dashed lines represent 10-channel QPSK signals with (green) high cubic distortion or (red) low cubic distortion.

Third-order distortion is usually the most important small-signal distortion in amplifiers, since it generates distortion products within and close to the band of interest. Third-order distortion is usually characterized by the (output) power at the third-order intercept point (OIP3), defined as that power where one of the two spurious products generated from the mixing of two nearby frequencies (tones) is equal in amplitude to one of the tones. This point is extrapolated from data acquired at low power levels. Spurious power resulting from third-order distortion varies as the cube of the signal power. Thus, the ACPR from third-order distortion varies as the square of the power, resulting in a slope of 2 on a logarithmic plot.

Any real amplifier can only supply a finite output voltage. For input signal levels beyond its capacity, the output becomes "clipped" or distorted. For multichannel digital signals with large peak-to-average power ratios, clipping will act first on the rare excursions to high amplitude. In the limit where the distribution of signal power is nearly Gaussian, the likelihood of a clipping event will be described by a complementary error function and, thus, the total power generated by clipping distortion will vary exponentially with input power, falling rapidly when the signal power is backed off from the clipping power by more than the peak-to-average power ratio.

As the signal becomes strongly clipped, the signal will become a square wave with a sin x/x spectrum. Thus, the adjacent-channel power will converge to a fixed value. For a Gaussian signal, the ACPR is approximately –12 dBc at input powers well beyond the clipping threshold.

Figure 4 shows the modeled adjacent-channel power for the same 10-channel QPSK signal in Fig. 3, using a simplified transfer curve including third-order distortion and hard clipping at a relative power of 0-dB input. Figure 4 demonstrates how the level of clipping distortion and third-order spurious power



6. The third-order intercept point is plotted versus operating frequency for generally similar GaAs HBT and MESFET amplifiers.

at an average signal power are strongly influenced by the type of signal employed. It can be seen that as the number of channels superimposed in the signal increases, the distortion behavior rapidly converges: four channels produce nearly the same peak-to-average ratio and distortion results as 10 channels. Note that the third-order distortion from a signal with high peak-average ratio is nearly equal to the two-tone third-order intermodulation, as predicted in ref. 3. This fortunate circumstance explains why a simple two-tone measurement is a useful guide to the likely value of third-order distortion for more complex signals.

Figure 5 shows the impact of varying the third-order distortion for a fixed input signal type. For input powers closer to the clipping power than the peak-to-average power ratio, distortion is dominated by clipping and third-order distortion behavior has no effect on the ACPR. However, as one input signal is "backed off" sufficiently, the third-order distortion takes over and sets a limit on the achievable ACPR.

Note that these simplified models do not take higher-order curvature of the transfer characteristic, or the detailed "shape" of the saturation behavior into account. Actual device behavior will differ from the predictions of the simple models, particularly at the intersection between the third-order and clipped characteristics, where higher-order terms can play a role and changes in the relative phase of the different contributions can cause fluctuations of several decibels in ACPR with modest changes in input power.

To fabricate amplifiers with good linear efficiency, component designers can employ processes and devices designed specifically for enhanced dynamic range, or seek circuit topologies which minimize the deleterious effects of distortion, or both. For example, at WJ Communications (San Jose, CA), the company's GaAs MESFETs have been optimized for low third-order distortion. This is accomplished by careful adjustment of the channel doping and geometry of the gate recess, ensuring that signal-dependent variations in device transconductance are almost perfectly nullified by the signal dependence of the device-output conductance. Most of the company's current high-dynamic range MESFETs are optimized for operation at zero gate bias ($I_{ds} = I_{dss}$), supporting operation from a single-voltage supply.

An advantage of MESFET technology is that the dominant nonlinear device elements are conductances, determined by doping concentration and electron mobility. Thus, the distortion behavior of MESFET amplifiers is relatively insensitive to variations in operating frequency and ambient temperature. But, MESFET nonlinear modeling is poorly understood compared to the nonlinear behavior of bipolar junction transistors (BJTs).

For high gain in limited semiconductor area, it is possible to use BJTs instead of FETs. The transconductance of a BJT is approximately proportional to the collector current: $I_c/(kT/q) = I_c/40$ at room temperature. For reasonable current densities, bipolar transistors can provide much higher transconductance per unit chip area than comparable MESFETs. The high gain supports the use of copious amounts of negative feedback while still preserving acceptable overall amplifier gain, and achieving low third-order distortion and good dynamic range. A Darlington configuration provides a low-impedance source (the first transistor) to drive the voltage gain stage (the emitter follower), decreasing sensitivity to parasitics, particularly the Miller capacitance. Heterojunction bipolar transistors (HBTs), with their heavily-doped base regions, also exhibit very low output conductance, making the output matching design simpler than for a conventional bipolar transistor circuit.

The feedback capacitance of a BJT is much larger than the corresponding capacitance in a MESFET, so the Miller effect magnifies the input capacitance in the second stage of a Darlington pair to produce significant gain rolloff. The presence of a significant nonlinear capacitance and the strong frequency dependence of the intrinsic gain cause the distortion behavior of BJT circuits to be more frequency dependent than MESFETs **(Fig. 6)**. Bipolar circuits require a resistor in the collector path, adding to power consumption. Finally, MESFETs tend to be more robust than BJTs when operating at high channel temperatures. The FET channel current decreases with increasing channel temperature due to reduced electron mobility, whereas BJT collector current increases with increasing temperature due to increased injection from the emitter, requiring careful design for good thermal stability. MRF

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2. E. Wesel, *Wireless Multimedia Communications,* Addison-Wesley, Boston, 1998, Ch. 4.
3. J. Pedro and N. de Carvalho, "On the Use of Multitone Techniques for Assessing RF Component's Intermodulation Distortion", *IEEE Transactions on Microwave Theory & Techniques,* Vol. MTT-47, 1999, p. 2393 (especially Fig. 5 therein).

Linear Amplifier Powers 80 W For MMDS Applications

This 2.5-to-2.7-GHz 80-W Class AB linear balanced amplifier is based on a quasi-enhancement-mode push-pull power device using surface-mount quadrature hybrids.

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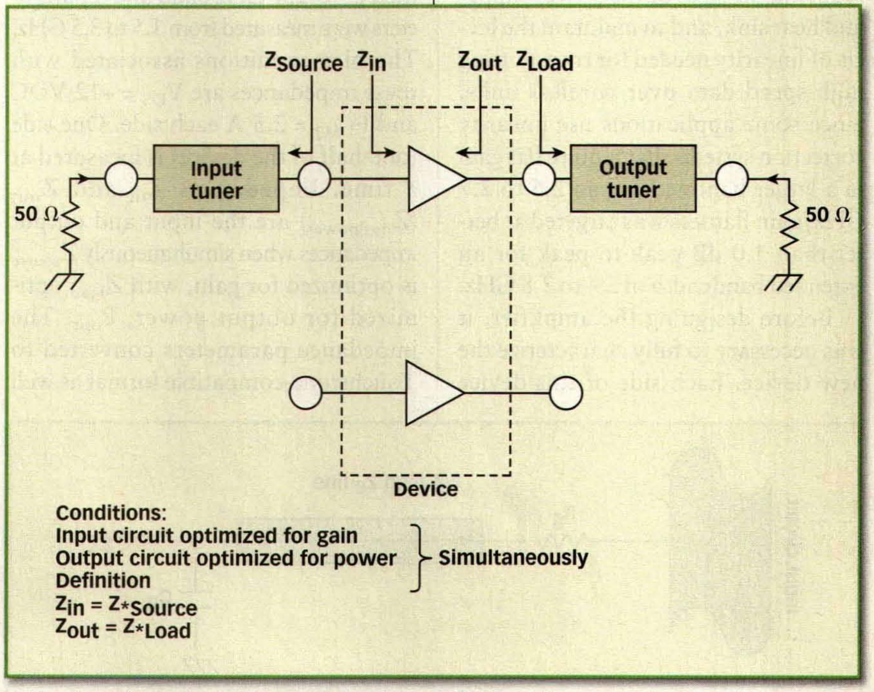
High power at high frequencies usually requires gallium-arsenide (GaAs) field-effect-transistor (FET) device technology, rather than the silicon laterally-diffused-metal-oxide-semiconductor (LDMOS) devices of lower-frequency cellular and personal communications services (PCS). One application that requires the use of GaAs FET RF power is in amplifiers for the multichannel multipoint distribution service (MMDS). To demonstrate, a linear Class AB amplifier was designed around a new quasi-enhancement-mode push-pull GaAs FET device. The amplifier, which is also suitable for wireless data/Internet technologies, achieves 80-W output power from 2.5 to 2.7 GHz with 12-dB linear gain and better than 1-dB gain flatness. It features 15-percent



1. The layout of a test system and system definitions for source/load-pull measurements are shown here.

power-added efficiency (PAE) when operating at 10-W average power for a wideband-code-division-multiple-access (WCDMA) signal, third-order intermodulation of −36 dBc for 10-W average power with a two-tone CW signal, and adjacent-channel power ratio (ACPR) of −44 dBc for 10-W average power with a WCDMA signal.

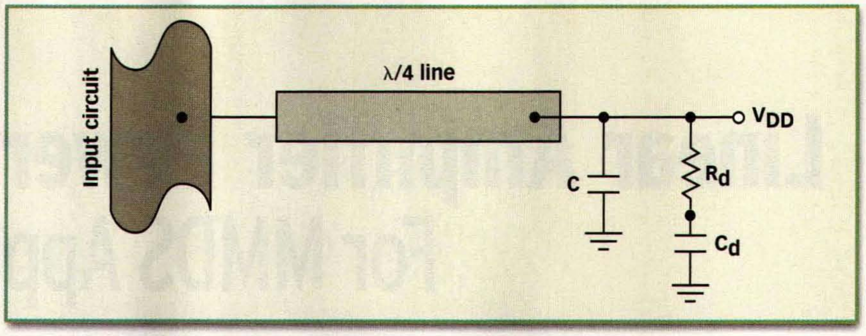
The new transistor, model FLL810IQ-3C, is a push-pull device using a pair of 40-W quasi enhancement-mode transistor chips optimized at +12 VDC for higher efficiency and good linearity at S-band. The device is rated for +49 -dBm typical output power at +12 VDC and 2.6 GHz, with 50-percent PAE under those operating conditions. The device offers a typical linear gain of 12 dB. The chips are matched for DC and RF operation and mounted with their input and output pre-matched circuits in an in-phase/quadrature (I/Q) push-pull package. Since it does not have any internal transversal connections between its two sides, the device doesn't require a virtual ground and it can be used in any amplifier configuration.

The design goals for the 80-W MMDS power amplifier (PA) include reducing the number of devices per amplifier, reducing the sizes of the power supply and heat sink, and to maintain the levels of linearity needed for transmitting high-speed data over wireless links. Since some applications use linearity correction systems that require flat gain in a larger bandwidth than 2.5 to 2.7 GHz, gain flatness was targeted at better than 1.0 dB peak-to-peak for an extended bandwidth of 2.4 to 2.8 GHz.

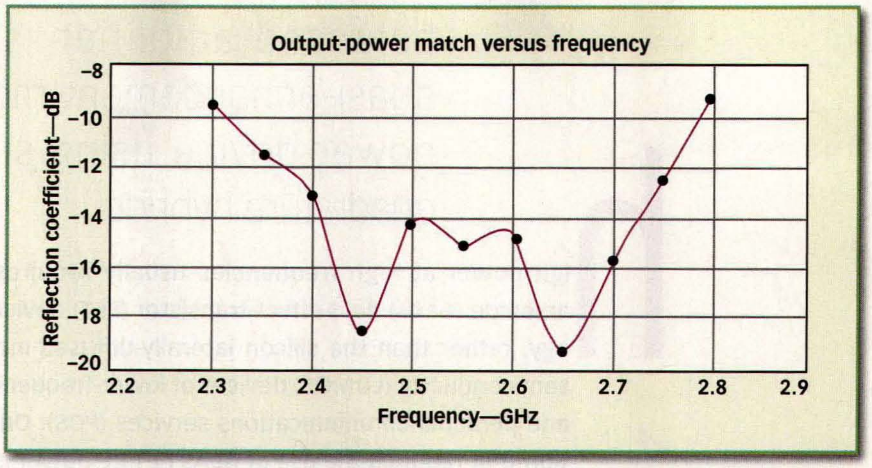
Before designing the amplifier, it was necessary to fully characterize the new device. Each side of this device was first characterized with a load/source-pull automated tuner system in the 2.3-to-2.8-GHz band and S-parameters were measured from 1.5 to 3.5 GHz. The bias conditions associated with these impedances are V_{DS} = +12 VDC and I_{DSQ} = 2.5 A each side. One side (one-half of the device) is measured at a time. Impedances Z_{in} and Z_{out} ($Z_{out|power}$) are the input and output impedances when simultaneously Z_{source} is optimized for gain, with Z_{load} optimized for output power, P_{out}. The impedance parameters converted to Touchstone-compatible format as well as S-parameters are available in the full-length version of the article appearing at the *Microwaves & RF* website at www.mwrf.com. The S-parameters are used to optimize the input circuit for gain and gain flatness and to analyze the amplifier stability. **Figure 1** provides the measurement methodology and the impedance definitions.

The FLL810-3C device allows designers to use any amplifier configuration. Two configurations are normally used with these devices: balanced and push-pull configurations. Both balanced and push-pull configurations[1] result in a similar basic performance when operating in any class of operation and in bandwidths smaller than one octave. Both approaches can be designed for similar linearity and efficiency performance as an amplifier for commercial application with relatively narrow bandwidth (10 percent). However, the balanced approach has several advantages over push-pull techniques:

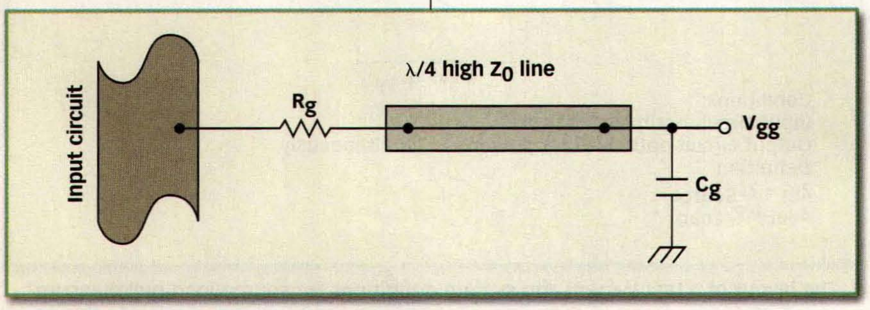
• Good external match supporting



3. This simple block diagram shows the drain-bias circuitry.



4. By matching the device output impedance to 50 Ω, while achieving good return loss, maximum output power can be achieved.



2. This simple block diagram shows the gate-bias circuitry.

a direct connection of the driver to the power stage.

- Higher isolation between the two sides of the device, resulting in better stability.
- Ease to design and integrate quadrature couplers for the amplifier.
- Better amplifier reliability, since if one side is damaged, the amplifier can still deliver one-quarter of its original rated power.

As a result, the balanced approach was used to design the 80-W MMDS amplifier using the FLL810-3C device.

Rogers 4350 circuit-board material from Rogers Corp. (Chandler, AZ) was selected for its acceptable loss, its suitable relative dielectric constant (3.48) for achieving small amplifier dimensions, and its relatively low cost. For the frequencies considered, several types of quadrature couplers may be used. Examples include two-branch printed couplers or surface-mounted hybrids. The latter was selected for its smaller size at the frequencies considered. For the 2.5-to-2.7-GHz band, the 90-deg. hybrid coupler achieved 0.12-dB insertion loss with 22-dB typical isolation and 1.10:1 typical VSWR.

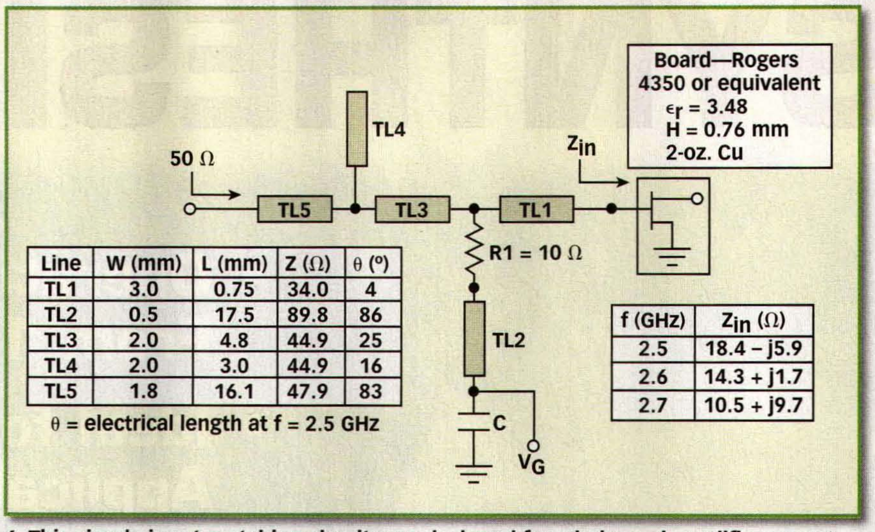
Each side of the push-pull device has its gate and drain-bias circuits. A 10-Ω gate resistor, R_g, is connected in series in each gate biasing circuit for gate protection and stability. Next is a quarter-wave length high impedance microstrip line short-circuited at its extremity by a capacitor with a series resonant frequency near 2.6 GHz. A high-impedance transmission line is used due to the low gate current. (See application note No. 010 from Fujitsu, "High-Power GaAs FET Device Bias Considerations," for more details.) **Figure 2** shows the gate-biasing circuit block diagram. Since the chips used in this device are DC matched, the two gate-bias circuits may be connected in parallel to the same gate-source voltage.

The amplifier drain-biasing circuit consists of a quarter-wave length microstrip line connected at one end to the output-matching circuit and at the opposite end short-circuited by a capacitor with a series resonant frequency of approximately 2.6 GHz. The quarter-wavelength microstrip line is low impedance since it has to carry up to 15-A maximum drain current when the device is in compression. In this design, a 2.5-mm-wide line is used to minimize the voltage drop in the line. **Figure 3** shows the drain-biasing circuit.

In each input and output circuit, a high-quality multilayer chip capacitor (ATC100A) with a series resonant frequency at 2.6 GHz is used as a DC blocking element. Its impedance at 2.6 GHz is very low in comparison to 50-Ω impedance and its insertion loss is practically negligible.

The amplifier circuit design was performed in several steps:

- The first step is to match the output impedance, Z_{out}, to 50 Ω while achieving a return loss of better than 15 dB in the bandwidth of interest. When this is achieved, the circuit should not be modified since no compromises should be made concerning the output-power performance. It is assumed that Z_{out} will not be significantly affected by a small change of source impedance, Z_{source}, during the gain and gain-flatness optimization procedures.
- The second step is to match the input impedance, Z_{in}, to 50 Ω in the band of interest while achieving a return loss of better than 15 dB. This defines the input-circuit configuration and the initial values of the circuit elements.

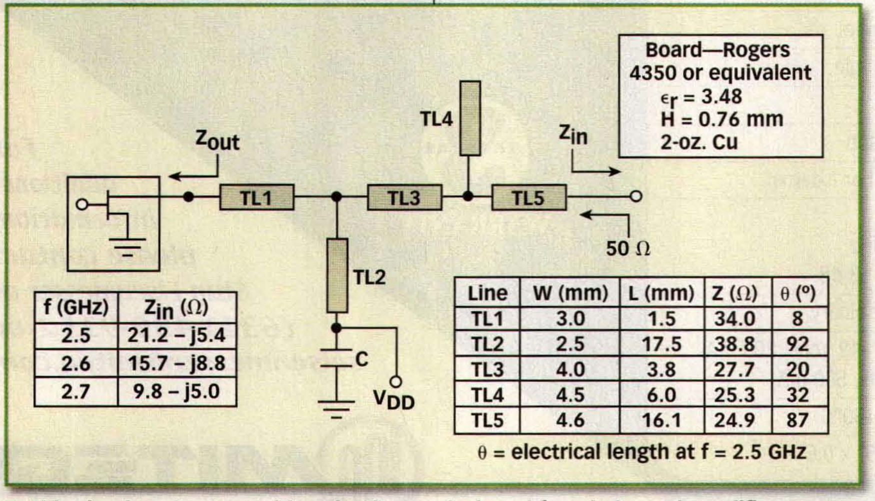


Line	W (mm)	L (mm)	Z (Ω)	θ (°)
TL1	3.0	0.75	34.0	4
TL2	0.5	17.5	89.8	86
TL3	2.0	4.8	44.9	25
TL4	2.0	3.0	44.9	16
TL5	1.8	16.1	47.9	83

θ = electrical length at f = 2.5 GHz

f (GHz)	Z_{in} (Ω)
2.5	18.4 – j5.9
2.6	14.3 + j1.7
2.7	10.5 + j9.7

6. This simple input-matching circuit was designed for a balanced amplifier operating from 2.5 to 2.7 GHz.



f (GHz)	Z_{in} (Ω)
2.5	21.2 – j5.4
2.6	15.7 – j8.8
2.7	9.8 – j5.0

Line	W (mm)	L (mm)	Z (Ω)	θ (°)
TL1	3.0	1.5	34.0	8
TL2	2.5	17.5	38.8	92
TL3	4.0	3.8	27.7	20
TL4	4.5	6.0	25.3	32
TL5	4.6	16.1	24.9	87

θ = electrical length at f = 2.5 GHz

5. This simple output-matching circuit was designed for a balanced amplifier operating from 2.5 to 2.7 GHz.

• The third step is to optimize the gain and gain flatness of the amplifier without the splitter/combiner. To do this, the S-parameters of the active device are used and only the circuit elements of the input circuit are optimized. It should be noted that the initial and final values of the input-circuit elements are very close due to the fact that the gain flatness when Z_{in} and Z_{out} are matched to 50 Ω is only 1.5 dB in the 2.4-to-2.8-GHz band.

• The fourth step is to analyze the amplifier stability for broadband application using the device S-parameters, without the input and output quadrature couplers.

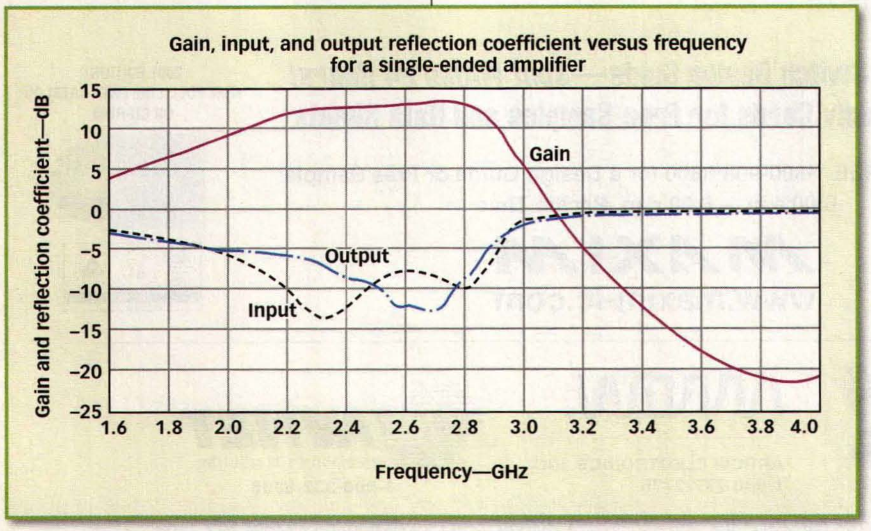
Finally, the global amplifier small-signal performance, gain, and external matching should be checked for the bandwidth of interest by using the device S-parameters as well as the S-parameters of the quadrature couplers.

Figure 4 shows the magnitude of the output-circuit reflection coefficient versus frequency to obtain maximum output power from 2.5 to 2.7 GHz. A return loss of better than 14 dB was obtained in the bandwidth of interest for high power. No gain consideration was made for the optimization of this circuit since the goal was to obtain maximum output power.

Comparing target and measured performance levels

PARAMETER	TARGET	MEASURED
Linear gain	11.5 dB	12.0 dB
Gain flatness (peak-to-peak) from 2.4 to 2.8 GHz	1.0 dB	0.6 dB
Saturated output power	+49 dBm	+49.4 dBm
Power-added efficiency at saturated output power	50 percent	50 percent
Two-tone CW IM3 for +40-dBm average output power	–35 dBc	–36.8 dBc
Two-tone WCDMA IM3 for +40-dBm average output power	–35 dBc	–35.5 dBc
ACPR at +40-dBm average output power	–40 dBc	–45 dBc
Power-added efficiency for WCDMA signal at +40-dBm average output power	15 percent	15 percent

The output-matching circuit is shown in **Fig. 5**. It consists of a transmission line TL1 that supports mounting the device without cutting its drain leads; a transmission line TL3 to bring the conductance, G, of the output admittance, Y = G + jX, to a value of G = 20 ms; a parallel open-circuited stub, TL4, to cancel the susceptance, X, of the admittance and obtain a real impedance, Z = 1/G = 50 Ω; and a transmission line, TL5 (usually used to transform the real impedance, Z, to 50 Ω), for a simple 50-Ω connection to the combiner port. A DC blocking capacitor is placed in a convenient location within transmission line TL5. It should be noted that the drain bias circuit (TL2 and C) does not have a matching function.



7. The gain, input reflection coefficient, and output reflection coeffcient of the single-ended amplifier design have been plotted from 1.6 to 4.0 GHz.

Figure 6 shows the input-matching circuit for optimal gain and gain flatness. A simplified circuit consists of transmission line TL1 which allows mounting the device without cutting its leads; transmission line TL3 to bring the conductance, G, of the input admittance, Y = G + jX, to a value of G = 40 ms; a parallel open-circuited stub, TL4, to cancel the susceptance, X, of the admittance and obtain a real impedance of Z = 1/G = 25 Ω; and a line, TL5, which is used to transform a real impedance of 25 Ω to a 50-Ω impedance.

It should be noted that the gate-bias circuit (R1, TL2, and C) has no matching function. The gate-bias circuit is connected as close as possible to the plane of the gate for stability and protection purposes.

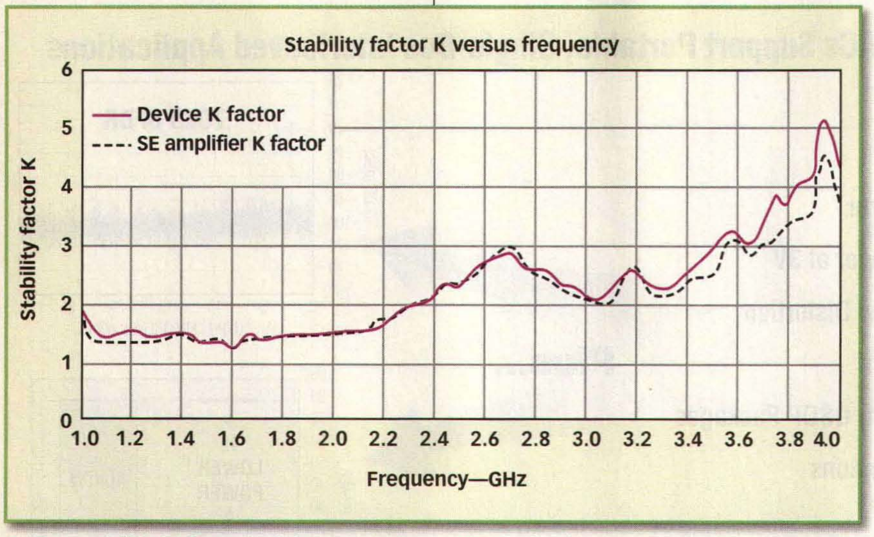
The complete single-ended amplifier circuit (one side of the amplifier) was analyzed using the device S-parameters measured from 1 to 4 GHz. The amplifier input circuit was optimized for gain flatness and gain from 2.4 to 2.8 GHz. Performance was also evaluated outside of that band for smooth amplitude rolloff **(Fig. 7)**.

The stability of the single-ended device at high frequencies was analyzed by using its S-parameters. The stability K-factor was calculated from 1 to 4 GHz and found to be greater than 1 **(Fig. 8)**. The gate resistors, drain R_d resistor/capacitor network, and decoupling capacitors should stabilize the amplifier at lower frequencies.

Two sides of the device were combined with two quadrature couplers and yielded peak-to-peak gain flatness of better than 0.4 dB and a minimum gain of 12.5 dB, which exceeding both target goals for these parameters.

The final amplifier circuit consists of two mirror images of the input- and output-matching circuits, one for each side. The complete set of decoupling capacitors recommended in application note AN-010 was connected in the gate- and drain-bias circuits. Since

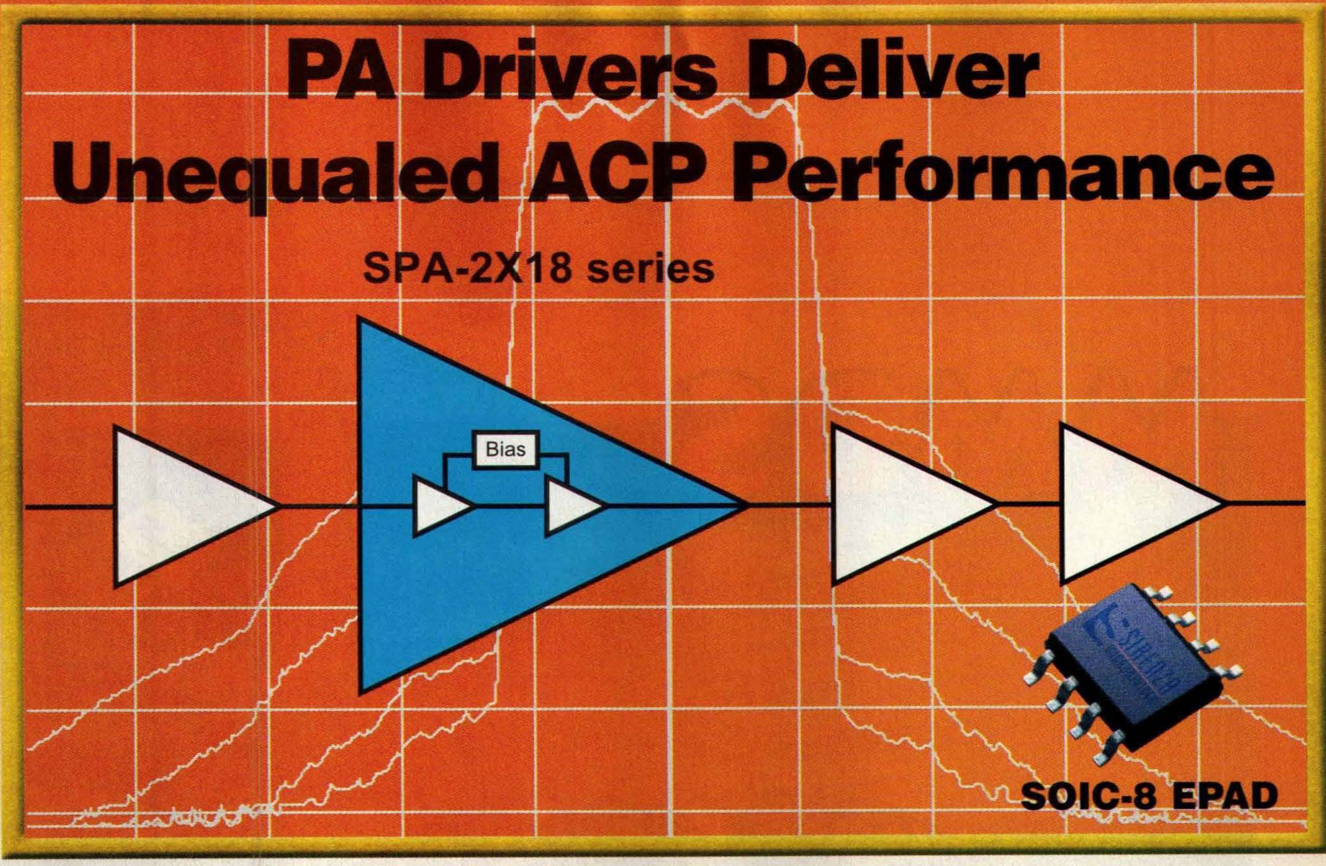
the S-parameters were measured only down to 1 GHz, it was not useful to model these capacitors with their via holes. The load resistor is a miniature 14-W power resistor mounted directly onto the amplifier housing in a hole in the circuit board. The dimensions of the amplifier board and device are 95 × 44 mm.



8. The stability (K factor) of the power GaAs FET device is compared with the stability of the single-ended amplifier design.

The best load/source-pull data are only accurate within 10 percent. Therefore, to obtain optimum results, fine tuning is necessary after amplifier assembly. The FLL810IQ-3C's internal prematching circuit helps to simplify the optimization of the external circuit. Once the tuning was found, consistent performance was observed among various devices. In fact, five devices were tested in the same amplifier tuned for output power with fixed tuning the check on the consistency of the device performance. The variation of output power across the five samples for an input power of +32 dBm is less than 0.6 dB peak-to-peak and the variation of output power for input power of +41 dBm is less than 1.2 dB peak-to-peak.

The data sheet for the FLL810IQ-3C GaAs FET provides the absolute maximum ratings for a flange temperature, T_f, of +25°C and gives the recommended drain-source voltage, maximum Igs,



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Competitor A	800-960	47	34	Not published	30.5	26	550	14.3
SPA-2318	1800-2200	47	30 (ACP tune) 28 (OIP3 tune)	21.5 dBm WCDMA 23.5 dBm IS-95	23	5	400	2.0
Competitor B	1800-2000	48	34	Not published	24	15	950	14.3
Competitor C	1500-2200	39	29.5	Not published	14	4.8	360	1.7

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and channel temperature, T_{ch}, for reliable operation. The channel temperature can be calculated from the device thermal resistance given in the data sheet using the formula:

$$T_{ch} = T_f + R_{th}P_{diss}$$

where:

T_{ch} = the channel temperature (in degrees Celsius),

T_{fl} = the flange temperature (in degrees Celsius), and

R_{th} = the device thermal resistance (in degrees Celsius/W or K/W).

The device power dissipated (in W), P_{diss}, can be found from:

$$P_{diss} = V_{ds}I_{ds} + P_{in} - P_{out}$$

where:

V_{ds} = the drain-to-source voltage (in V),

I_{ds} = the drain-to-source current (in A),

P_{in} = the input power (in W), and

P_{out} = the output power (in W).

When $P_{in} = P_{out} = 0$ W (no RF signal), $I_{ds} = I_{dsq}$ and $P_{diss} = V_{ds}I_{dsq}$,

where:

I_{dsq} = the quiescent current.

The thermal resistance given in the data sheet is only value under-the-test conditions presented in the data sheet, since GaAs material thermal conductivity is a strong function of temperature. This means that the data sheet gives R_{th1} for a defined set of conditions: T_{f1}, T_{ch1}, or P_{diss1}. For a different set of conditions (T_{f2}, T_{ch2}, or P_{diss2}), the new R_{th2} can be calculated by using the methodology presented in ref. 2. For defined operating conditions, T_{ch} can be defined and the device mean time to failure (MTTF) can be calculated by using a curve of process MTTF versus channel temperature for the FLL810IQ-3C (available in the extended version of this article on the *Microwaves & RF* website at www.mwrf.com, which also provides thermal data).

The final amplifier met or exceeded all of the target performance goals. The results presented in **the table** represent the average of four amplifiers. All four were tuned for optimum two-tone WCDMA IM3 performance at V_{ds} = +12 VDC and I_{dsq} = 5 A, at 2.6 GHz and +40-dBm average output power. MRF

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1. J. Shumaker, R. Basset, and A. Skuratov, "High-Power GaAs FET Amplifiers: Push-Pull versus Balanced Configurations. Example W-CDMA (2.11-2.17 GHz), 150-W Amplifiers," Wireless Symposium & Exhibition, February 12-16, 2001.
2. R. Basset, "Understanding Thermal Basics For Microwave Power Devices," *Microwaves & RF*, October 2000, pp. 101-110.

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Market	Frequency (GHz)	Output Power P1dB (dBm)	Saturated Power (dBm)	Gain (dB)	PAE (@ Psat %)	Part Number
UNII & HiperLAN	5.0 - 6.0	26	+29	18	38	HMC406MS8G
		25	+29	15	28	HMC407MS8G
		29	+32	20	25	HMC408LP3
		23	+26	20	35	HMC415LP3
Wireless Local Loop	3.0 - 4.0	27	+30	21	45	HMC327MS8G
		28.5	+32	25	25	HMC409LP3
Cellular	1.5 - 2.3	27	+30	20	45	HMC413QS16G
MMDS	2.1 - 3.2	27	+30	20	32	HMC414MS8G

HMC327MS8G

- 3.0 - 4.0 GHz
- Output P1dB: 27 dBm
- Saturated Power: +30 dBm
- Gain: 21 dB

HMC406MS8G

- 5.0 - 6.0 GHz
- Output P1dB: 26 dBm
- Saturated Power: +29 dBm
- Gain: 18 dB

HMC413QS16G

- 1.5 - 2.3 GHz
- Output P1dB: 27 dBm
- Saturated Power: +30 dBm
- Gain: 20 dB

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DESIGN

Interpret And Apply EVM To RF System Design

Understanding the definitions of EVM and relating it to design parameters, such as LO phase noise, is critical for RF communication system development.

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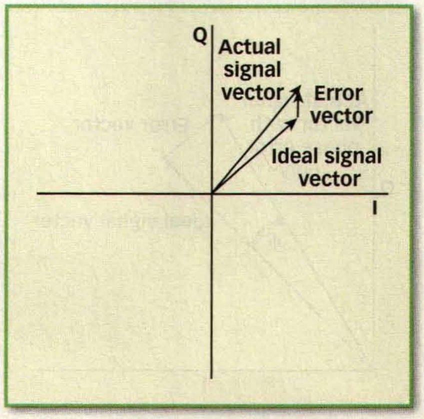
Error vector magnitude (EVM) has been a measure of modulation accuracy for several years. The concept of EVM is simple—it is the magnitude of difference between the ideal modulation vector and the actual modulation vector. Since the actual modulation vector has been changed by nonidealities (nonlinearities and phase noise) in the system, EVM represents a measurement of the errors that are introduced by the system.

For engineers unfamiliar with EVM, the defining equations are less than intuitive. Performing some basic geometric and trigonometric analysis goes a long way toward gaining an understanding of EVM. Once this understanding has been established, one can begin to ask questions such as "How does local-oscillator (LO) phase noise contribute to EVM?" and "How do EVM contributors add together?"



1. This drawing shows the signal and error vectors.

Figure 1 shows the concept of EVM. The magnitude of the error vector is a measurement of how far from ideal the actual transmitted vector has been degraded. If the EVM is large enough, it is easy to see that the vector may be incorrectly interpreted as a different symbol than what was intended. System designers impose bounds on EVM in order to keep transmission and reception errors within acceptable limits.

EVM In EDGE Systems

Engineers who are experienced in designing Global System for Mobile Communications (GSM) and, in particular, Enhanced Data Rates for GSM Evolution (EDGE) systems will recognize the definition for calculating EVM according to the European Telecommunications Standards Institute (ETSI). This is described in GSM 05.05 annex G.[1] It states that a symbol's EVM shall be computed at the symbol times during the useful part of the transmitted burst. It begins by modeling the actual transmitted symbol vectors as Z_k, which are

the actual transmitted complex vectors transmitted at time instant k, one symbol apart.

$$Z_k = \{C_0 + C_1 * [S_k + E_k]\} * W^k \quad (1)$$

where:

k = the discrete time in the EDGE system; k = n/270.833 kHz for EDGE,

C_0 = a constant origin offset representing quadrature modulator imbalance during the burst. Quadrature modulator DC offsets are one key contributor to this constant.

C_1 = a complex constant that represents the arbitrary phase and output power of the transmitter (Tx) during a burst. This constant may change from burst to burst, but is constant during any burst.

S_k = the ideal transmitted signal that is observed through the measuring filter and sampled at time k.

E_k = the residual error vector on sample S_k.

$W = e^{dr + jda}$ accounts for a frequency and amplitude offset across the burst. This would result in a fixed phase change of "da" radians per symbol which one could imagine comes from main injection synthesizer frequency error over the burst, and a fixed amplitude change of "dr" nepers per symbol, which might come from things such as average amplitude droop over the burst. Remember that these are fixed numbers for any particular burst, but will also change from burst to burst.

Defining Neper

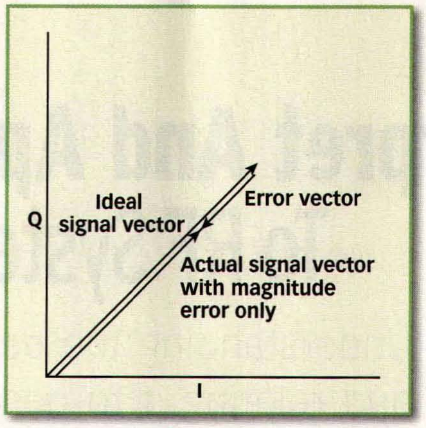
A neper is a dimensionless measurement of the ratio of two amplitudes. One can talk about changes in amplitude of a signal in Nepers without necessarily knowing what the units of the signal are. The neper is often used to express voltage and current ratios, where decibels (dB) is often used to express power ratios.

Specifically, a value in Nepers (Np) is given by:

$$Np = ln(x_1/x_2) \quad (2)$$

where:

x_1 and x_2 = the two amplitudes.

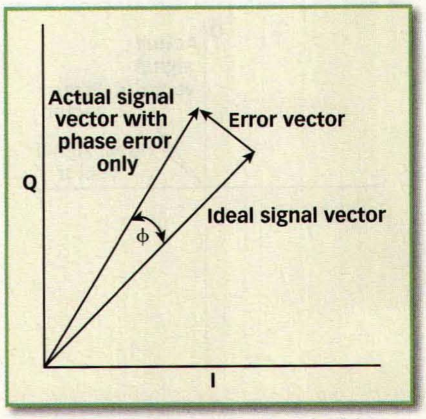


2. The signal with magnitude error only can be seen here. This is a closeup view of an ideal signal vector and an actual signal vector.

All of these burst-dependent constants—C_0, C_1, and W—must first be calculated for each burst so that they minimize root-mean-square (RMS) EVM per burst, and then they are to be used for computing the individual vector errors E_k. This math is computationally intense and is most conveniently performed on a computer. Rearranging the expression for actual transmitted vectors Z_k provides an expression for E_k:

$$E_k = \{[Z_k * W^{-k} - C_0]/C_1\} - S_k \quad (3)$$

Although this looks like a complex expression, it can be interpreted simply. The error vector, E_k, is the difference between the actual transmitted vector Z_k (after being normalized for the frequency and amplitude errors that are constant across the burst), and the ideal transmitted signal vector (i.e., if there



3. This figure shows the signal with phase error only.

was not any amplitude droop or frequency error across the burst, $W^{-k} = 1$ and no EVM). If the modulator did not have any quadrature imbalance, C_0 equals 0. C1 simply normalizes the actual signal against the Tx fixed output power and phase. If all of these hold, the expression is simplified to:

$E_k = Z_k - S_k$ = actual transmitted vector – ideal transmitted vector

This is illustrated in Fig. 1.

Once the individual error vectors, E_k, have been computed, the individual EVM or general EVM can be computed. GSM 05.05 version (Ref. 1) describes this as:

$$EVM_k = \left[|E_k|^2 / (|SK|^2 / -K)\right]^{0.5} \quad (4)$$

which is the error vector length relative to the root average energy of the useful part of the burst. This is often multiplied by 100 to express EVM as a percent. Also note that there is an 05.05 expression for RMS EVM across one burst:

$$RMS\ EVM = \left(|E_k|^2 / - |S_k|^2\right)^{0.5} \quad (5)$$

Equation 5 represents the root energy of all error vectors in the useful part of the burst—relative to the root energy of the ideal signal vectors in the useful part of the burst.

Ideal Signal

Next, consider an ideal signal that develops a magnitude-only error (no phase error). This is shown with a closeup view of an ideal signal vector and an actual signal vector as shown in **Fig. 2.** Suppose that the ideal signal vector passes through an amplifier that causes the amplifier to compress, which caused an error in the magnitude of the ideal transmitted signal.

Note that the ideal and actual vectors in Fig. 2 are offset for clarity. In reality, they would both have origins exactly at (0,0).

Let M_{dB} = the magnitude of error in decibels.

First, convert the magnitude error in decibels to a numeric value in order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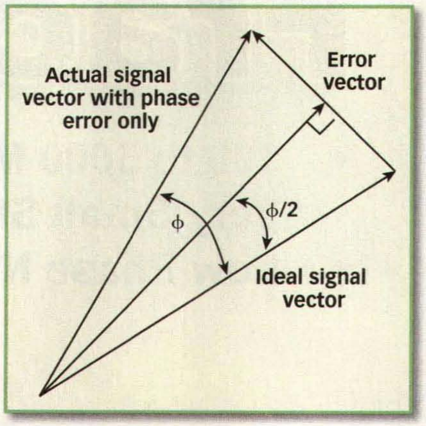
to find the length of the transmitted vector.

$$M_{dB}\ (\text{numeric}) = 10^{MdB/20}$$

Realize that if the error were 0 dB (an ideal vector), the vector magnitude would be 1 (i.e., normalizing to the ideal vector). Next, find the difference between the ideal vector length and the actual transmitted vector length (this is the error vector).

$$|E_k| = |Z_k - S_k| = \left|10^{MdB/20} - 1\right| \quad (6)$$



4. This is a closeup view of Fig. 3, which shows the signal with phase error only. One can zoom in on the triangle made by the vectors, and bisect the original triangle into two right triangles.

In order to find the EVM, one needs to divide the EVM by the ideal vector magnitude, and multiply by 100 to figure out the percentage.

$$EVM = \left[|E_k|/|S_k|\right] * 100\% \quad (7)$$

A short example will now be given. Suppose the ideal signal drives an amplifier into compression, which causes a 0.3-dB magnitude error to the ideal signal. What would the EVM be? Applying Eq. 6 provides:

$$|E_k| = \left|10^{0.3/20} - 1\right| = 0.035 \quad (8)$$

Applying Eq. 7 provides:

$$EVM = \left[|E_k|/|S_k|\right] * 100\% = 3.5\% \quad (9)$$

That is, a 0.3-dB magnitude error has caused a 3.5-percent EVM. Predicting EVM that is based on headroom to amplifier P1dB depends greatly on the shape of the amplifier's output-power versus input-power curve and on the statistics of the signal vector (peak-to-average ratio and more).

EVM From Phase Error

The EVM that is from phase error for a single symbol is less obvious than the magnitude error question, but with another closeup diagram and some trigonometry, the result can be understood **(Fig. 3)**. Suppose now that the ideal signal vector passes through an RF mixer where the

LO phase noise imparts a phase error onto a single symbol (**Fig. 4**). One can zoom in on the triangle that is made by the vectors, and bisect the original triangle into two right triangles. Then use the definition of sine to figure out the length of the error vector.

$$|E_k| = 2 * sin(\phi/2) = 2 * sin(0.04/2) = 0.04 \quad (12)$$

$$EVM = [|E_k|/|S_k|] * 100\% = [0.04/1] * 100\% = 4.0\% \quad (13)$$

Let Ø = phase error in degrees. From Fig. 4, one can use the definition of sine to figure out what half of the error vector length is:

$$|E_k|/2 = sin(\phi/2)$$

$$|E_k| = 2 * sin(\phi/2) \quad (10)$$

Once an expression for the magnitude of the error vector (Eq. 10) is obtained, one can apply Eq. 7 in order to find the EVM. To illustrate this, a short example will be used. Suppose an LO imparts a phase error of 2.3 deg. to an ideal signal vector as it passes through a mixer. The question is what the resulting EVM will be. First, convert to radians (not necessary, but done here for clarity).

$$\phi = 2.3 * 3.14_{rad}/180^\circ = 0.04 \text{ radians} \quad (11)$$

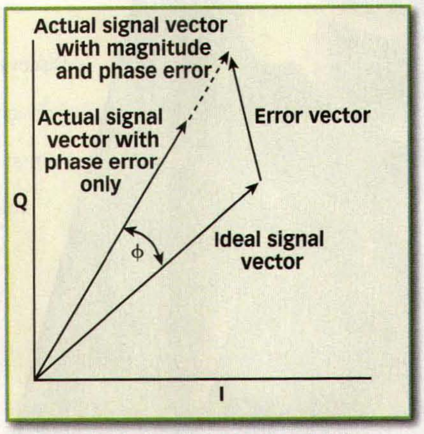
Apply Eq. 10:

(See Eq. 12 above)

Apply Eq. 7:

(See Eq. 13 above)

In brief, a 2.3-deg. phase error has caused a 4-percent EVM. It is interesting to compare the GSM 5-deg. RMS phase-error budget (from GSM 05.05) to EDGE 7-percent RMS EVM budget. Using the same calculations as



5. An actual signal vector containing a magnitude and phase error can be seen here.

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before, a 5-deg. phase error, which would pass the GSM specification, causes an 8.4-percent EVM, which fails the EDGE EVM specification. Equation 10 can be used when estimating LO phase-noise contributions to system RMS EVM. If the RMS phase error of the LO is known, one can plug it into Eq. 10 in order to predict the RMS EVM contribution due to LO RMS phase error.

EVM that is from magnitude and phase error for a single symbol is similar to, but slightly more complicated than the case of phase error alone. First, consider the diagram in **Fig. 5**, now with an actual signal vector containing a magnitude and phase error.

In **Fig. 6**, one can observe a closeup view of the vectors. Let Ø represent the phase error in radians, and let M be equal to the magnitude error (numeric, not in decibels). From Fig. 6, one can invoke the law of cosines in order to solve for the length of the error vector in terms of the known phase error, Ø, along with the known magnitude error, M.

(See Eq. 14 below)

One can check the sanity of this expression by checking it against the magnitude-only (let Ø = 0) as well as the phase-error-only (let M = 0) work that was done earlier.

For the case of magnitude error only, let Ø = 0, then cos Ø = 1, and

(See Eq. 15 below)

Now rearrange what is inside the square-root sign:

(See Eq. 16 below)

Recognize this as the difference of two numbers, squared.

$$|E_k| = \left\{\left[1 - (1 + M)\right]^2\right\}^{0.5} \qquad (17)$$

Now the square-root sign cancels with the exponent (one needs to only

$$|E_k| = \left[(1 + M)^2 + 1^2 - 2(1 + M)\cos\phi\right]^{0.5} \qquad (14)$$

$$|E_k| = \left[(1 + M)^2 + 1^2 - 2(1 + M)\right]^{0.5} \qquad (15)$$

$$|E_k| = \left[1^2 - 2(1 + M) + (1 + M)^2\right]^{0.5} \qquad (16)$$

$$|E_k| = \left[(1 + 0)^2 + 1^2 - 2(1 + 0)\cos\phi\right]^{0.5} \qquad (20)$$

care about the magnitude of M).

$$|E_k| = 1 - (1 + |M|) \qquad (18)$$

The 1s cancel, leaving the following expression:

$$|E_k| = |M| \qquad (19)$$

$$|E_k| = [4(1 - \cos\phi)/2]^{0.5} = 2[(1 - \cos\phi)/2]^{0.5} \qquad (23)$$

$$|E_k| = \left[(1 + M_k)^2 + 1^2 - 2(1 + M_k)\cos\phi_k\right]^{0.5} \qquad (25)$$

$$|E_{total}| = \left[\Sigma(1 + M_k)^2 + 1^2 - 2(1 + M_k)\cos\phi_k\right]^{0.5} \qquad (27)$$

This makes sense, since if the error was magnitude only, the error vector equals the magnitude error.

For the case of phase error only, let M = 0, then:

(See Eq. 20 on page 91)

Simplify this:

$$|E_k| = (1 + 1 - 2\cos\phi)^{0.5}$$

$$|E_k| = (2 - 2\cos\phi)^{0.5} \qquad (21)$$

Multiply what is inside the square-root sign by 1 (= 2/2):

$$|E_k| = [2(2 - 2\cos\phi)/2]^{0.5} \qquad (22)$$

Rearrange:

(See Eq. 23 above)

Use the half-angle identity to rewrite what is left.

$$|E_k| = 2 * \sin(\phi/2) \qquad (24)$$

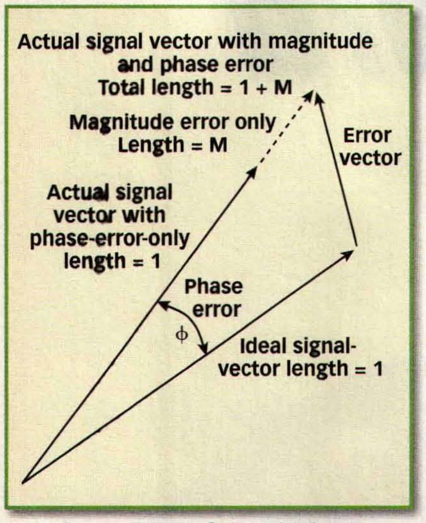
This is the same result as Eq. 7, as it should be, since this was the phase-error-only case.

One can imagine that EVM sources add together in an RMS sum fashion in the way that noise sources add. On the other hand, one might suppose that for EVM, with its magnitude and phase errors, it might be more accurate to RMS sum the magnitude errors first, then RMS sum the phase errors, and finally convert the magnitude and phase sums to EVM using the formula derived in Eq. 14.

Method To Use

Opinions seem to differ on which method to use, or if it matters at all. The following text discusses the two methods.

• Method 1—convert each individual source of magnitude and phase



6. A closeup view of Fig. 6 is shown above—signal with magnitude and phase error.

error into EVM, and then RMS sum the individual EVMs from all sources.

Consider K sources of magnitude and phase error in that $M_1,Ø_1$ are magnitude and phase error of first source; $M_2,Ø_2$ are magnitude and phase error of second source; while $M_k,Ø_k$ are magnitude and phase error of K^{th} source.

The EVM for source number k (k = 1...K) is known from the result of Eq. 14.

(See Eq. 25 on page 92)

The total EVM using this method would be the RMS sum of all |Ek|.

$$\left|E_{total}\right| = \left(\Sigma\left|E_k\right|^2\right)^{0.5} \qquad (26)$$

Or, written as a single expression:

(See Eq. 27 on page 92)

• Method 2—RMS sum all sources of magnitude error for a total magnitude error. Then RMS sum all sources of phase error for a total phase error. Finally, convert the total magnitude and phase errors to EVM.

Consider again, K sources of magnitude and phase error where $M_1,Ø_1$ are magnitude and phase error of the first source; $M_2,Ø_2$ are magnitude and phase error of the second source; and $M_k,Ø_k$ are the magnitude and phase error of K^{th} source.

First, RMS sum all magnitude errors for total magnitude error:

$$M_{total} = \left(\Sigma\left|M_k\right|^2\right)^{0.5} \qquad (28)$$

Next, RMS sum all phase errors for total phase error:

$$\phi_{total} = \left(\Sigma\left|\phi_k\right|^2\right)^{0.5} \qquad (29)$$

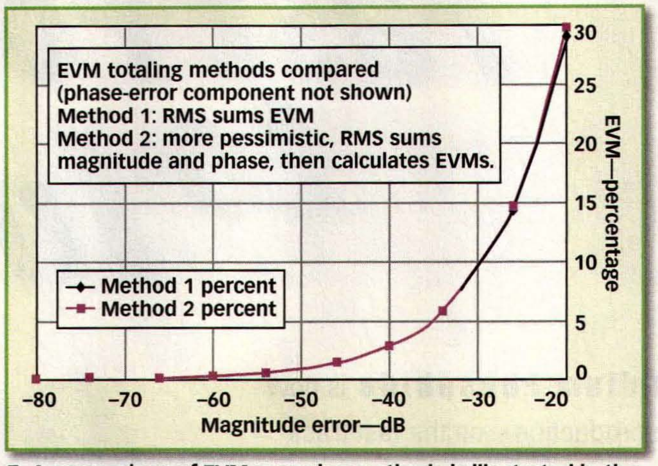
Now convert the total magnitude and phase error to EVM:

(See Eq. 30 below)

Or, write as a single expression:

(See Eq. 31 below)

The expressions for EVM totals from methods 1 and 2 are not equal. However, for small values of EVM (that is, an EVM of less than 0.5 percent), they are equivalent down to 0.001 percent. Method 2 is more pessimistic, but it does not differ from method 1 by more than 0.1 percent—even for EVM totals of approximately 10 percent. An Excel spreadsheet **(Fig. 7)** was generated in order to compare the two methods for four EVM contributors of equal magnitude and phase errors. The results of that spreadsheet will be shown.



7. A comparison of EVM summing methods is illustrated in the graph above.

Opinions differ on which method is better. Results for the two methods are similar but method 2 is more pessimistic. Method 2 might be chosen since it is more conservative.

The phase errors were such that they caused the same amount of EVM as the magnitude errors.

Not every EVM contributor in an RF system should be summed in RMS fashion. For example, if two elements, such as SAW filters, are cascaded, their EVM contributions may add directly. This can be understood by visualizing that if the in-band ripple (which causes magnitude errors on an FM-type signal such as GSM) of the filters is exactly the same, this is similar to creating an overall filter with twice the in-band ripple of a single filter alone. This could double the EVM. It must be accounted for in an EVM budget analysis. On a related note, if the SAW filter responses can be such that the peak in one, is the valley in another, the cascaded ripple may be lower than the individual ripple, which may actually improve EVM.

Finally, applying geometric and trigonometric analysis to the definition of EVM can broaden one's understanding of the definitions. Mathematical modeling of various stages in communication systems can further this understanding. MRF

$$\left|E_{total}\right| = \left[\left(1 + M_{total}\right)^2 + 1^2 - 2\left(1 + M_{total}\right)\cos\phi_{total}\right]^{0.5} \qquad (30)$$

$$\left|E_{total}\right| = \left\{\left[1 + \left(\Sigma\left|M_k\right|^2\right)^{0.5}\right]^2 + 1^2 - 2\left[1 + \left(\Sigma\left|M_k\right|^2\right)^{0.5}\right]\cos\left[\left(\Sigma\left|\phi_k\right|^2\right)^{0.5}\right]\right\}^{0.5} \qquad (31)$$

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Uncover Bluetooth Packet Errors

The combination of a Bluetooth test set and an oscilloscope can be used to trigger on Bluetooth packet errors as part of the process of maximizing data throughput.

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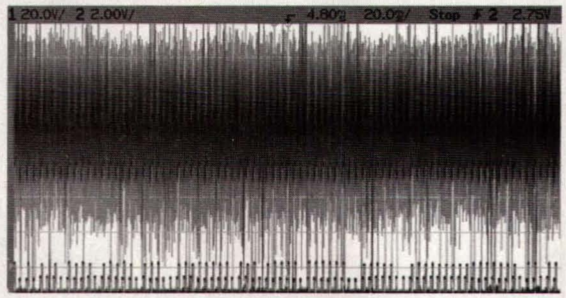
Optimization of data throughput is one of the key tasks in creating highly integrated Bluetooth solutions. Bluetooth promises remote control and data access without wires, using low-cost radio technology at 2.4 GHz. But to maximize the efficiency of a Bluetooth system, it is first necessary to optimize the Bluetooth data throughput by examining Bluetooth data packets and errors as a way of optimizing data-transfer rates in a Bluetooth-enabled design.

Bluetooth, of course, is the wireless connectivity standard developed a few years ago by Ericsson and several other communications and computer companies, including IBM, Intel, and Nokia. A Bluetooth network consists of a master device and one or more slave devices. Up to seven slave devices can be active in one time in a miniature Bluetooth network, known as a piconet, although additional devices can be in an idle state. The idle devices can become active by assuming one of the "positions" of an active device.

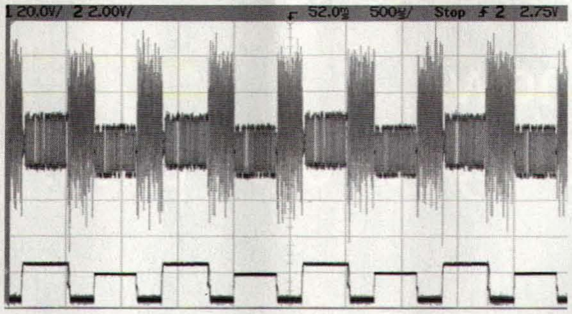
Bluetooth operates in the 2.4-GHz industrial-scientific-medical (ISM) band, using 79 channels in most countries, each with a bandwidth of 1 MHz, although some countries have adopted variations of this scheme (notably Japan, France, and Spain) in which only 23 channels are used.

Bluetooth data are transmitted at a raw data rate of 1 Mb/s with Gaussian frequency-shift-keying (GFSK) modulation employing a bandwidth-time (BT) product of 0.5. The transmission scheme is based on time-division-duplex (TDD) methods, using 625-μs timeslots. Transmissions from the master device must begin in an even-numbered timeslot, while transmissions from the slave devices must begin in an odd-numbered timeslot. Transmissions are allowed to span multiple time slots, to a maximum of five



1. During a 200-ms acquisition time, 150 BlueRF packets were acquired (top trace) with the test set, along with the power envelope, which is a signal that is proportional to the transmitted or received power (the lower trace).

timeslots. This allows available bandwidth to be divided between uplink and downlink traffic (asymmetrical) at various rates. When Bluetooth is operating with purely symmetrical data transfers, the maximum data rate is 185.6 kb/s. But when operating in an asymmetric fashion, the maximum data rate is 721 kb/s. Bluetooth operates with slow frequency hopping, at a hop rate of 166 hops/s or one hop per timeslot.



2. A couple of the originally acquired BlueRF packets, magnified to examine the details of the packets, are shown here.

Packet Data

Bluetooth packetized serial data are often referred to as BlueRF. Although the name may infer a high-frequency signal, BlueRF refers to the demodulated data. One measure that indicates the overall operational health of a Bluetooth system is packet-error ratio (PER), with maximum system performance only being achieved when packet errors are minimized. Engineers may be intrigued by the fact that as the transmitter/receiver (Tx/Rx) is being designed and tradeoffs are being decided with respect to Tx and Rx parameters, it is difficult to ultimately track the final real-world data-transfer rates back to these tradeoffs. For example, even though a Bluetooth Rx may be designed with high sensitivity and good co-channel and adjacent-channel performance, the final data-transfer rates will be affected by the interaction of these specifications, along with the various variables of the actual working environment. PER is a single measure, where the variables of Tx, Rx, and the operating environment coalesce.

The first task is to "see" a packet error. Once that is accomplished, design engineers can actually start to understand the root causes of the error. A proven method for capturing and viewing BlueRF signals is with an oscilloscope. The instrument can be used for capturing and correlating BlueRF signals together, along with other associated control and trace lines. Once an effective trigger is determined, it can also be used for triggering other instruments, including a spectrum analyzer.

Most engineers will start their Bluetooth design by purchasing an evaluation board that they ultimately intend to incorporate in their design from one or more of the Bluetooth chip or module vendors. Depending on the specific product design of the Bluetooth evaluation board, engineers may or may not have access to all of the signals that are needed to trigger the oscilloscope when a dropped packet occurs. Using a Bluetooth test set, such as the model E1852A from Agilent Technologies (Santa Rosa, CA), provides engineers with an easy way to access the BlueRF, even when the design is integrated to a level where it may be difficult or impossible to directly probe. The Bluetooth test set allows designers to establish a link with a Bluetooth device under test (DUT), and to measure parameters including power, bit errors, and packet errors. Some Bluetooth test sets also include signal outputs that can be used to trigger test equipment such as an oscilloscope. In the case of the model E1852A, there are outputs for the BlueRF, signal power, slot clock (625-µs-spaced pulses) and a data-received pulse. These particular signals can be used as effective triggers.

Packet errors can be examined in multiple test modes as well as in the normal operational modes of a Bluetooth system. The following examples use a Bluetooth system that is in loopback test mode, where expectations are that the same information will be routed from the Tx out to the Rx, with the Rx sending the same packet back again. The packet type and payload length will determine the fundamental timing of the interactions, which will become key information when looking for errors. If a system does not support loopback test mode, the same principles that are used to find packet errors will apply for normal mode, with a poll from the Tx replacing the sent packet.

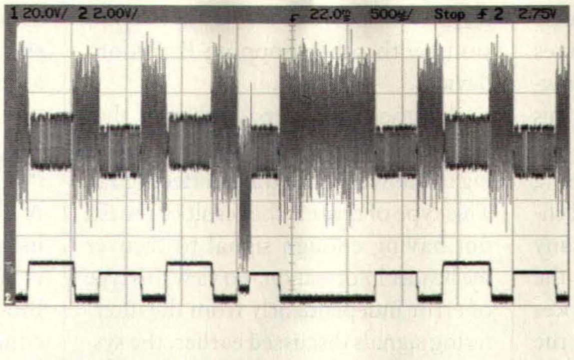
Acquiring Signals

Once this link is established and the appropriate signals—the BlueRF and other key control signals in the design—are being probed, a single acquisition can be taken to examine the signals, thereby initiating a baseline for the signals that are being investigated. Taking this single acquisition with a deep-memory oscilloscope enables the long time frames that are associated with the serial transfer of data to be captured at a resolution that supports a thorough post examination. Now the acquired information will be scanned for anomalies, which, in this case, is dropped packets.

As **Fig. 1** shows, 150 BlueRF packets have been acquired during 200 ms (the top trace), along with the power envelope, which is a signal that is proportional to the transmitted or received power (the lower trace). Since so much data has been collected, it is difficult to analyze it at this level. However, if the power envelope is examined carefully, breaks can be seen in the pattern of the lower power signal being sent from the Tx and the higher power response from the Rx. These breaks are packet errors and in this one capture, there are more than 15 packet errors. This corresponds to a packet-error rate of 10 percent, and the effect on the system is a 10-percent decrease in the maximum system throughput.

Figure 2 shows a couple of the originally acquired BlueRF packets, magnified to examine the details of the packets. This is a good example of a nor-

mal exchange sequence without any packet errors, with the top trace representing the BlueRF and the bottom delineating the power envelope. In this particular loopback test, DH1 packets are carrying a full payload. This can be confirmed by counting the number of 625-μs slots that the packet occupies. In this case, it is easy to verify that the data packet is one slot long. It is important to note that if fewer bits are transferred in the payload, the packet will occupy less than a full slot. This becomes important later, when the "length" of time that a packet occupies can be used as a means of triggering the oscilloscope.



3. A trace of the BlueRF and power envelope is shown here. In this case, it is enlarged around an area of the original acquisition suspected of an error.

Now, as design engineers begin looking for errors, conditions under which they occur can be determined. This entails sifting through the original acquired data for information that does not conform to this repetitive interaction expected in loopback mode—that is, tracking down packet errors. One-way packet errors occur when the received packet is corrupted to a point where it is unrecognizable by the Rx, with no response being sent as a consequence. The original captured data is panned through, arriving at one of the areas identified earlier as a location of a possible problem and then zooming in to uncover more detail.

Figure 3 again shows a trace of the BlueRF and power envelope, in this case enlarged around an area of the original acquisition that is suspected of an error. While there is a power pulse associated with the transmitted signal, there is no pulse associated with a signal being sent back from the Rx. Noting the missing power pulse can provide an easy way to trigger the oscilloscope, capturing the dropped-packet events and associated signals for further examination. Some typical causes become evident when the details of the original 200-ms acquisition are examined more closely, and the areas where dropped pulses/packets can be identified are explored. In the case of Fig. 3, an interfering signal has distorted the transmitted packet.

PACKET ERRORS

This type of error happens in a noisy environment, when multiple sources are transmitting in the ISM band. However, Bluetooth architects anticipated this kind of problem, which is one of the reasons for the pseudorandom hopping sequence that is used in Bluetooth technology. Bluetooth devices hop into any one of 79 discrete frequencies in the ISM band. This allows Bluetooth devices to communicate even if some of the frequencies are occupied by other more powerful sources. At this point, a spectrum analyzer could also be triggered, using the trigger output created by the oscilloscope when a dropped packet is detected to determine which frequencies or channels are being most affected by interference. This interference might be at a single frequency, indicating a "fixed" frequency source, such as a microwave oven. Or, the errors may be distributed throughout the 79 channels, indicating a more distributed interference, such as random collisions with other hopping Bluetooth devices.

Another type of packet error that can occur is one created when reducing the power of the transmitted signal. This type of error is the result of the Rx not having enough signal to recover the header information. To view this type of error independently from the interfering signals discussed earlier, the system needs to be RF shielded. This can be accomplished with an RF-shielded enclosure for the DUT, and providing a direct-wired RF connection from the test set to the Bluetooth device under test. When viewing this type of packet error on an oscilloscope, there will typically be no noticeable difference between a packet sent by the Tx that elicits a response from the Rx or one that is dropped. Examining this type of error can provide design engineers with insight into the performance of their systems as the distance between devices increases, or to optimize their systems to operate at the lowest power possible.

Optimizing Performance

A Bluetooth system can only achieve its maximum performance when packet errors are minimized. By using the Bluetooth test set and oscilloscope combination to trigger on these packet errors, the packets that are not recovered can be examined. Now, with a way to look at packet errors, engineers may be able to find ways to recover some of these lost packets, which is a major advantage since the individual bits in the packet cannot be recovered if the whole packet has been lost. Knowing how to capture these events will enable those engineers designing Bluetooth systems to construct error-correction schemes and optimize their designs. MRF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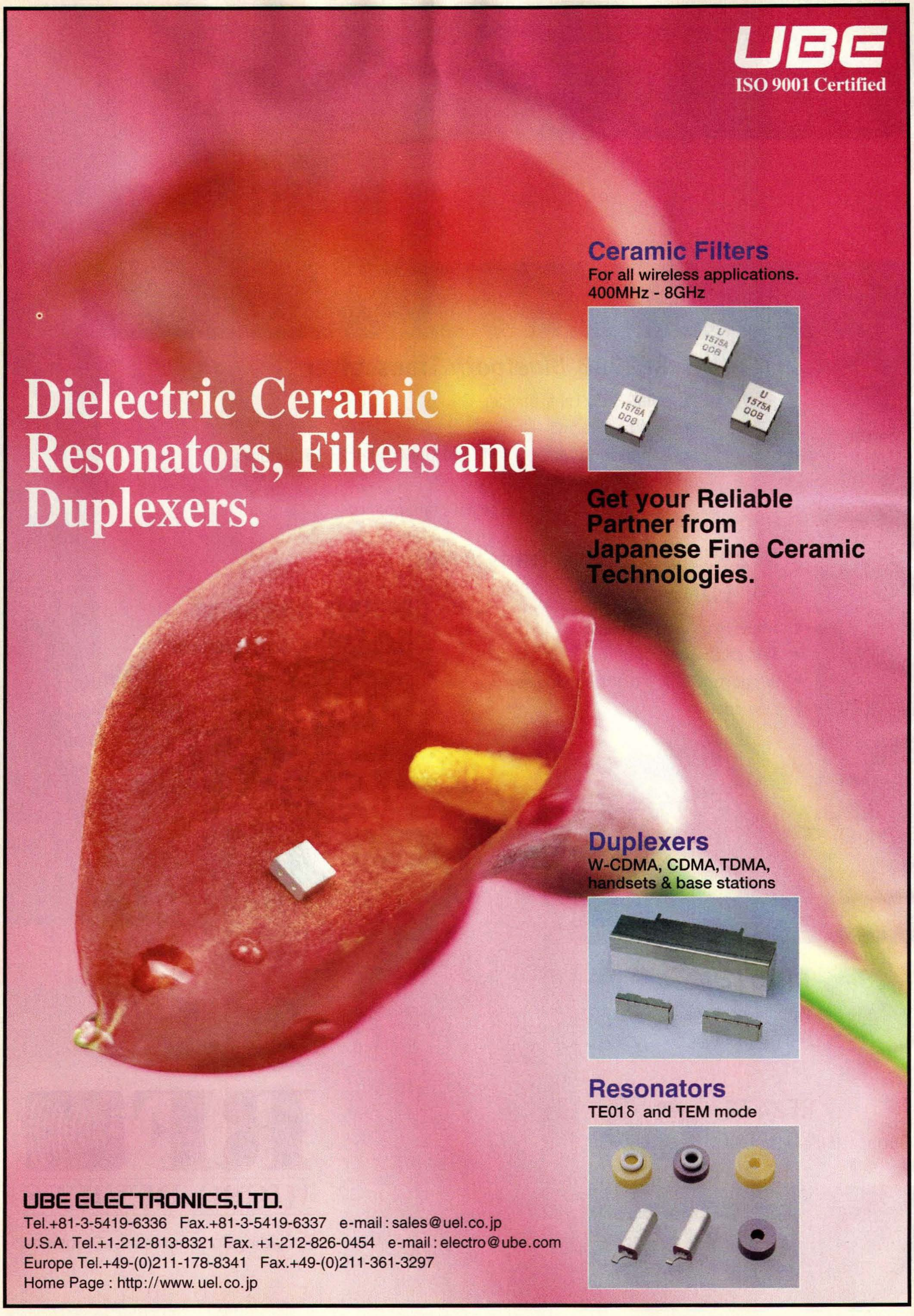
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BLUETOOTH

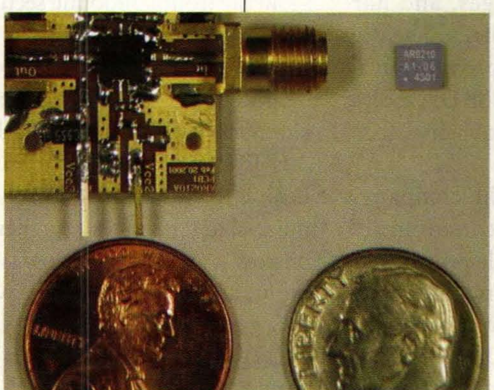
Amplifier Drives Bluetooth And Wireless Data

This efficient amplifier provides better than +25-dBm output power in the wireless data and Bluetooth bands of 902 to 928 MHz and 2400 to 2500 MHz.

JACK BROWNE
Publisher/Editor

Wireless data applications come in all shapes and frequency ranges. Two of the more popular bands, the ISM bands of 902 to 928 MHz and 2400 to 2500 MHz, can be served with a single high-efficiency amplifier from Araftek, Inc. (Fremont, CA). The model AR0210 features 45-percent efficiency with +25-dBm typical output power over the Bluetooth range of 2400 to 2500 MHz.

The model AR0210 amplifier **(see figure)** is actually usable from 500 to 2500 MHz. It is fabricated with a GaAs HBT process and is designed with maximum RF input-power levels to +10 dBm, and typically draws 165-mA current at +3.5 VDC. The idle current is only 39 mA when operating with a supply of +3.5 VDC and an RF input level of −30 dBm.



The AR0210 is a high-efficiency GaAs HBT amplifier that is designed for wireless data applications at 902 to 928 MHz and 2.4 to 2.5 GHz.

For applications from 902 to 928 MHz, the AR0210 amplifier typically achieves maximum output power of +26 dBm with typical PAE of 58 percent. Forward isolation is typically 35 dB, while second and third harmonics are typically −40 dBc. All other spurious levels are typically −50 dBc. Over this frequency range, the amplifier offers adjustable gain from 0 to 28 dB, under analog control through 0- to +6-VDC applied voltage.

For Bluetooth applications from 2400 to 2500 MHz, the AR0210 amplifier typically achieves maximum output power of +25 dBm with 45-percent typical PAE. Forward isolation is typically 25 dB, while second and third harmonics are typically −45 and −40 dBc, respectively. All other spurious levels are typically −50 dBc. The gain can be varied from 0 to 30 dB through 0- to +3.5-VDC control voltages.

The AR0210 amplifier is suitable for wireless data terminals, WLANs, Bluetooth systems, and any portable wireless equipment requiring up to 250-mW transmit power at the antenna port. It is supplied in a compact 4 × 4-mm QFN-16L plastic package, and is usable at operating temperatures from −40 to +85°C. The company also offers the model AR0211 amplifier with an integral power detector and similar RF performance, but for a gain-control range of 0 to 22 dB. **Araftek, Inc., 40990 Encyclopedia Circle, Fremont, CA 94538; (510) 580-2500 ext. 203, FAX: (510) 580-2508, Internet: www.araftek.com.** MRF

Take A Crash Course In Phase Shifters

SHIFTING THE PHASE of a sinusoidal-type signal seems straightforward theoretically, but using phase-shift networks in an application requires an understanding of the subtleties of packaged phase-shift circuits. This information is available in an application note, "PHASE SHIFTERS, Electronic & Mechanical; Analog & Digital, 200 kHz to 3 GHz," from Merrimac Industries, Inc. (West Caldwell, NJ). The five-page note begins by describing the various types offered by the company—mechanical, electronic, and digital—and then describes the limitations and trade-offs associated with each type.

For example, the limited number of variable capacitor styles reduces the range of phase shift for mechanical shifters, but for electronic devices, the theoretical phase-shift range is unlimited since they use varactors instead of manually adjusted capacitors. Insertion loss is a parameter that designers must understand and deal with at two levels. With electronic shifters, the phase shift versus control-voltage curve can be linearized using a multisection approach where only the linear portion of each phase-shift section is used. But this increases insertion loss. Shifters having no inductors are phase linear, but maximum phase shift is limited to 90°. To obtain greater phase shift, stages can be added but at the cost of greater insertion loss.

In the section of the note called Parameter Definitions, the company points out that the phase shift versus frequency curve (group delay) is not constant. That is, the value and slope of the phase shift varies as a function of frequency. Also, phase stability is a function of temperature, meaning that the insertion phase-deviation at any frequency is a function of temperature. This can result in a variation of ±4° over a temperature range of −60 to 100°C, depending on the input-phase angle. Other parts of this section provide information on insertion-loss variation versus control voltage, settling time, input power, and additional topics important to specifying the correct type of phase shifter for the application.

The application note is available for downloading from the company's website.

Merrimac Industries, Inc., 41 Fairfield Pl., West Caldwell, NJ 07006; (973) 575-1300, FAX: (973) 575-0531, Internet: www.merrimacind.com.

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A scope measures voltage, current, and, with a few accessories, can measure instantaneous power, floating voltages, and harmonics present on the supply voltage.

Scopes Make Ideal Switching Power-Supply Design Analysis Tools

SWITCHING POWER SUPPLIES are standard in virtually all electronic systems and with switching frequencies in the hundreds of kilohertz and heading higher, the oscilloscope becomes a prime tool for making power measurements. A scope measures voltage, current, and, with a few accessories, can measure instantaneous power, floating voltages, and harmonics present on the supply voltage. The way to make these measurements is provided in an application note entitled, "TDS3000B DPO Solves Today's Power Measurement Problems," from Tektronix (Beaverton, OR).

The note is based around the company's TDS3000B series of DPOs equipped with an FFT Application Module. This module provides the scope with the ability to make and display the relative magnitude of the harmonics to the fundamental frequency. Also part of the 3000B's optional accessory package are a high-voltage differential probe and a current probe. Together, these two probes support an instantaneous power measurement because they can be used to make a floating voltage measurement simultaneously with a current measurement. A section of the note is devoted to the subject of deskewing of the voltage and current probes, which is critical to the correct measurement of instantaneous power. Deskewing equalizes the timing delay between the voltage and current probes to produce an accurate reading.

In the troubleshooting section of the note, information is provided on the viewing of modulation effects in a current-mode control loop. Transient capture is another important measurement tool in power-supply design, and the 3000B offers two functions, Roll Mode and Peak Detect, which permit the capture of slow noise-signal changes together with glitches as narrow as 1 ns, even at slow sweep speeds. Harmonic analysis is performed with the FFT module which offers spectrum-analyzer-type frequency components. The note can be downloaded from the Tektronix website.

Tektronix, Inc., Beaverton, OR; Internet: www.tek.com/Resources For You/App Notes/Oscilloscopes.

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cover story

Low-Power IF IC Digitizes 300 MHz

This flexible IF digitizer IC can capture signal bandwidths as wide as 270 kHz with better than 90-dB dynamic range.

Selection of a wireless receiver (Rx) architecture has long involved a compromise between the relative merits of superheterodyne Rxs versus direct-conversion Rxs. While the former offers wide dynamic range, it is generally more complex and requires more power than lower-performance direct-conversion Rxs.

Fortunately, the model AD9874intermediate-frequency (IF) digitizing integrated circuit (IC) from Analog Devices (Wilmington, MA) swings that choice in favor of the narrowband superheterodyne approach by integrating all of the function blocks needed for IF-to-digital conversion except the voltage-controlled oscillator (VCO). The AD9874 provides the wide dynamic range associated with superheterodyne Rxs, but with the simplicity and power consumption that are associated with direct-conversion Rxs.

In next-generation cellular base stations, smaller size, lower power, and lower costs are driving reasons for redesign. Global System for Mobile Communications (GSM) base-station designers are seeking ways to achieve these advancements while maintaining the overall integrity of the system. The AD9874 takes a novel approach to partitioning the IF strip of a narrowband radio Rx that helps design-

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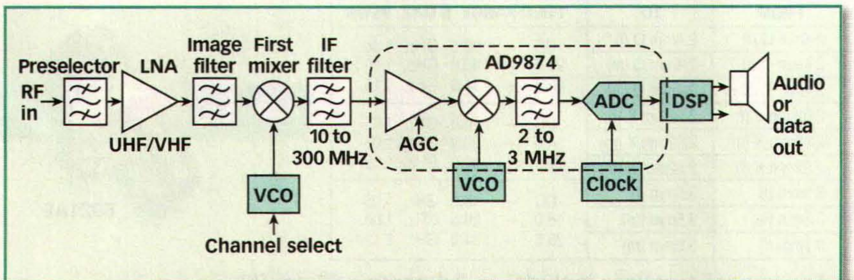
Analog Devices, Inc., 604 Woburn St., Wilmington, MA 01887; (781) 937-1175, FAX: (781) 937-1027, e-mail: paul.hendriks@analog.com, Internet: www.analog.com.

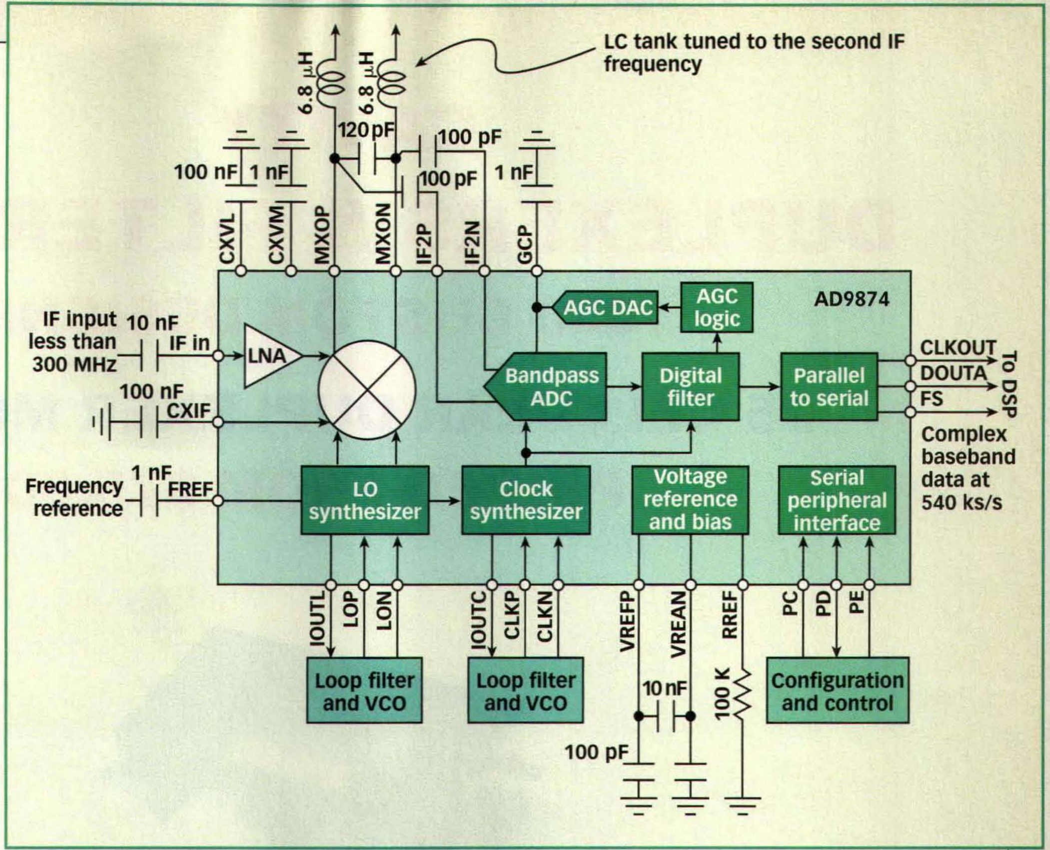


1. This dual-conversion superheterodyne Rx architecture has been designed with the AD9874 IF digitizing subsystem.

ers meet their power, size, and cost goals while achieving high dynamic range.

The superheterodyne Rx, an example of which is illustrated in **Fig. 1**, is a popular architecture known for high dynamic range. The implementation shown avoids quadrature analog downconversion to baseband by using filtering together with multiple analog downconversion operations, thereby side stepping troublesome issues associated with quadrature analog downconversion and low-frequency analog signal processing. Constructing this Rx is complicated by the need to find a good frequency plan and the need to deal with a multitude of sub-blocks.

The AD9874 simplifies the



2. This "inside" look at the AD9874 IF digitizing subsystem is configured for a 26-MHz clock frequency and set for maximum output signal bandwidth.

design problem by integrating the bulk of the radio's back end into a single IC whose performance is guaranteed by the manufacturer. Furthermore, since the input IF range of the AD9874 is very broad and since all modulation-specific functions are delegated to a digital signal processor (DSP), the AD9874 can be used to construct a reconfigurable radio platform.

Comparing the AD9874 and the AD9870

PARAMETER	AD9874	AD9870
Maximum clock frequency	26 MHz	18 MHz
Maximum signal bandwidth	270 kHz	150 kHz
Noise figure	10 dB	12 dB
Third-order intercept point	-1 dBm	-1 dBm
Instantaneous dynamic range	90 dB	80 dB
Current consumption	22 mA	45 mA

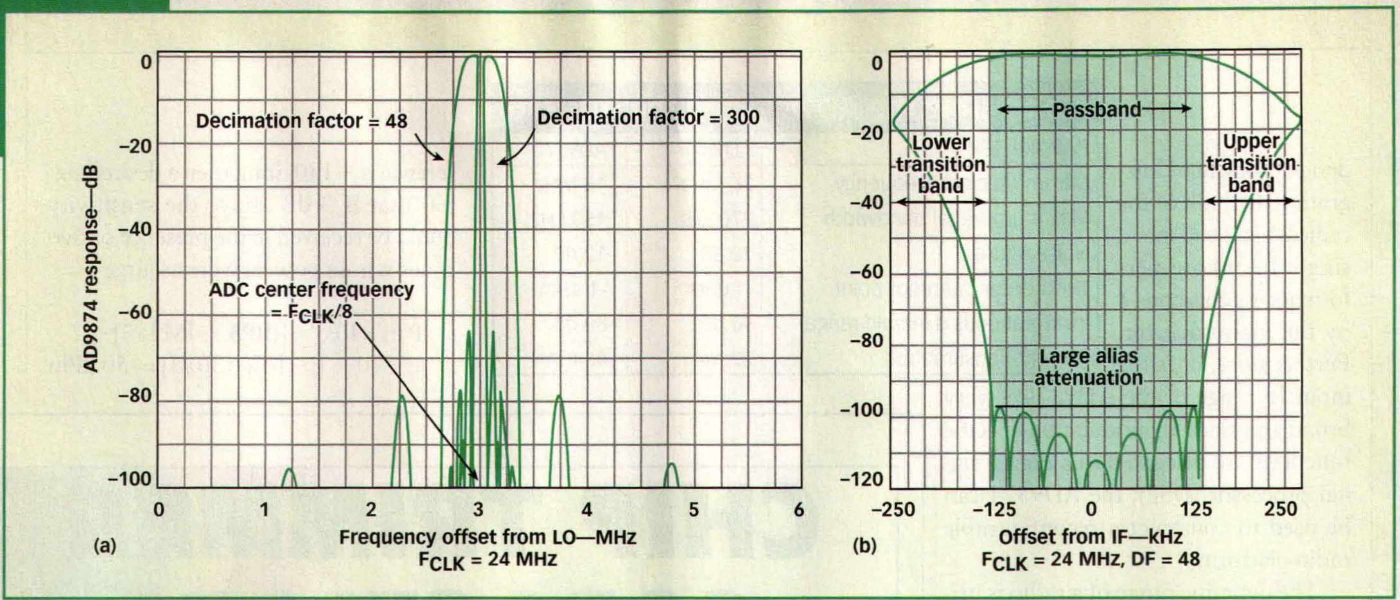
The dynamic range of a radio is primarily a function of three key parameters. The first is its noise figure, which is the ratio of the input-referred noise power to the thermal noise limit (-174 dBm/Hz, or $0.5\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$ in a 50-Ω system). A sensitive Rx such as that found in a base station might have a noise figure of only a few decibels. Provided with the noise figure and signal bandwidth, a system designer can calculate the signal-to-noise ratio (SNR) as a function of the signal power, and vice-versa. For example, if an Rx with a noise figure of 3 dB and a signal bandwidth of 10 kHz needs 6 dB SNR to detect an input with acceptable fidelity or bit-error rate (BER), the sensitivity limit of the Rx is:

$$-174\ \mathrm{dBm} + 10\log_{10}(10 \times 10^3) + 3 + 6 = -125\ \mathrm{dBm}$$

or approximately 0.12-μV RMS in a 50-Ω system.

The second key parameter affecting Rx dynamic range is its linearity. The most common linearity specification in an analog receive chain is the input third-order intercept point (IP3). The input IP3 is the extrapolated input power needed by two tones to produce a third-order intermodulation (IM) product equal in level (amplitude) to those two tones. From the input IP3 specification and the characteristic 3-dB/dB slope of third-order distortion, a system designer can calculate the tolerable blocker amplitude for two-tone interference. For example, if an Rx with an input IP3 of -10 dBm experiences a 3-dB loss in sensitivity when the effective in-band interference is -130 dBm, then a desired signal that is 3 dB above the sensitivity could be received in the presence of two tones whose powers were as large as:

$$P\ in = IIP3 - [(IIP3 - IM3)/3]$$
$$= -10 - [(-10 + 130)/3] = -50\ \mathrm{dBm}$$



3. The typical performance of the AD9874 IF digitizing subsystem is compared with its predecessor, the AD9870, including (a) filtering characterisitics and with (b) output data oversampled by 2X (the signal bandwidth is one-half the output data rate).

In the context of a system employing a hard-limiting element such as an analog-to-digital converter (ADC), another consideration is the input level which results in clipping. The instantaneous dynamic range (IDR) of an Rx is the ratio between the power of a single interference signal which causes sensitivity degradation and the sensitivity limit, and is measured with all automatic-gain-control (AGC) circuits set at minimum attenuation. The IDR expresses an Rx's ability to cope with a single large interference signal and is as, if not more, important a measure of Rx linearity as input IP3 is when an ADC is part of the signal chain. It is more likely to have a single large interferer than to have two large interferers with comparable powers situated at frequencies that create an on-channel distortion product.

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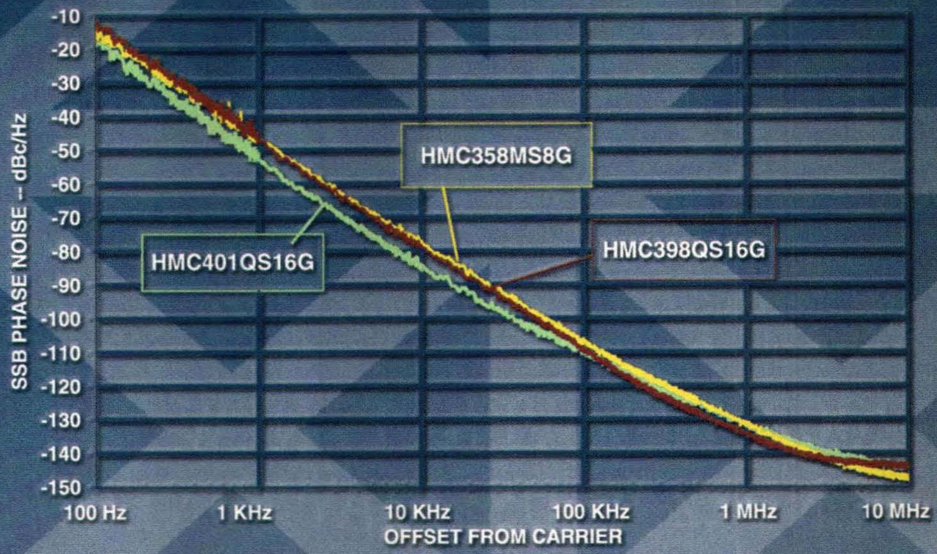
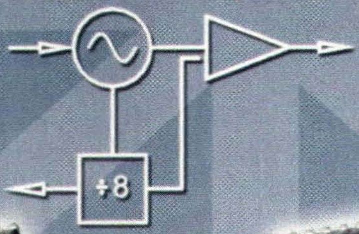


MMIC VCO w/ Buffer Amplifier

KU-Band MMIC VCO w/ Divide-by-8



HMC358MS8G

- 5.6 - 6.8 GHz
- Phase Noise: -105 dBc/Hz
- Pout: > +10 dBm

HMC398QS16G

- 14.0 - 15.0 GHz
- Phase Noise: -110 dBc/Hz
- Pout: > +6 dBm

HMC401QS16G

- 13.2 - 13.5 GHz
- Phase Noise: -110 dBc/Hz
- Pout: > -8 dBm

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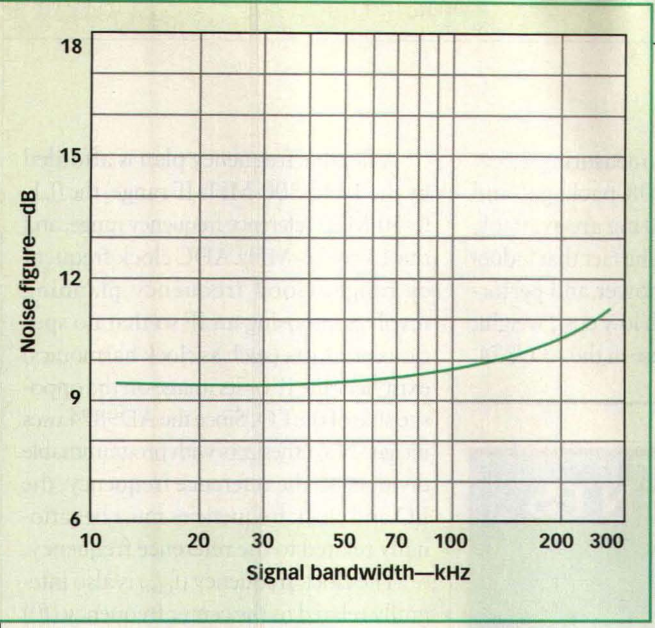

4. This plot shows the trade-offs between noise figure and bandwidth (with an IF of 109 MHz and 26-MHz clock rate).

As shown in **Fig. 2**, the AD9874 accepts an IF input up to 300 MHz and outputs complex baseband [in-phase/quadrature (I/Q)] data. The data stream is serial, consisting of 16/24-b I and Q data followed by an optional 16-b AGC/status word. To facilitate a trade-off between pin usage and decode complexity, the AD9874 supports several 1-, 2-, and 3-wire serial formats.

The primary blocks within the IC are a low-noise amplifier (LNA), a mixer, an ADC, and a digital filter. The LNA presents a 360-Ω resistance to the IF in pin and drives the doubly-balanced mixer with an amplified version of the IF in input. The mixer downconverts this signal to a second IF which is then digitized by the bandpass ADC. A digital filter mixes the bandpass data down to baseband and also filters the output down to the desired bandwidth. Finally, the parallel-to-serial converter outputs the data in the desired serial format.

Secondary blocks within the IC include local-oscillator (LO) and clock synthesizers, a serial-peripheral-interface (SPI) block, as well as AGC and bias circuitry. The LO and clock synthesizers lock the LO and clock signals produced by external LO and clock VCOs to the frequency reference applied to the FREF pin. Alternatively, either synthesizer can be placed in standby and LO/clock signals applied directly to the IC. The SPI port provides a three-wire interface to the status and control registers within the IC. The AGC circuitry prevents clipping in the ADC with large signals by reducing the gain in the signal path.

A unique aspect of the AD9874 is its use of inductors as signal-processing elements. Inductors are much-maligned components but are the key to the AD9874's ability to accurately digitize a bandpass signal while consuming only a small amount of power. A comparison of the AD9874 with the earlier model AD9870 serves as evidence of this point **(see table)**. The AD9870 uses switched-capacitor technology to realize a fully-integrated bandpass ADC, whereas the AD9874 takes advantage of external inductors in the construction of its ADC. As a result of this architectural difference, the AD9874 is able to achieve 10 dB more instantaneous dynamic range while simul-

taneously cutting the power consumption in half.

Modern inductors are small, rugged, and inexpensive. Since tolerances of ±10 percent are easily accommodated by the on-chip tuning circuitry, precision components are not needed. Suitable surface-mount inductors measuring 2.5 × 2.0 × 1.6 mm (a 1008 package) and weighing less than 100 mg are available for $0.08 in quantity. The fact that inductors yield significant power and performance advantages with low cost, weight, and size justifies their use in the AD9874.

A flexible frequency plan is afforded by the 10-to-300-MHz IF range, the 0.1-to-50-MHz reference frequency range, and the 13-to-26-MHz ADC clock frequency range. Good frequency planning involves choosing an IF so that no spurious products (such as clock harmonics) exist near the IF or its image on the opposite side of the LO. Since the AD9874 uses integer-N synthesizers with programmable dividers on the reference frequency, the LO and clock frequencies must be rationally related to the reference frequency.

The clock frequency (f_{CLK}) is also integrally related to the center frequency (f0) of the ADC as well as to the output data rate. In the case of the center frequency this ratio is fixed at 1:8 (i.e. $f0 = f_{CLK}/8$) to be compatible with the first-generation AD9870. This integer relationship is used because it simplifies the internal hardware which performs quadrature digital downconversion.

In a digital radio, the clock frequency is typically a multiple of the symbol rate. Since 13 MHz is a standard frequency within a GSM Rx (13 MHz is 48 times the 270.833-kHz baud rate), the AD9874 is designed to accept either a 13- or 26-MHz clock. For the sake of universality, any frequency between these limits is also allowed.

In the AD9874, the output signal bandwidth is equal to half of the output data rate, which is in turn equal to the clock frequency divided by the decimation factor (DF). DF can be one of 48 n or 60 n, where n ranges from 1 to 16. For example, with a clock frequency of 24 MHz the signal bandwidth can be as low as 12.5 kHz or up to 250 kHz. **Figure 3a** illustrates the filtering characteristics of the IC with DF = 48 and DF = 300, while **Fig. 3b** shows an expanded view of the passband for DF = 48. The steep cutoff and massive attenuation of signals folding into the ±125-kHz passband are hallmarks of digital filtering. Since the AD9874 uses finite-impulse-response (FIR) filters, these desirable attenuation characteristics are provided by filters which have perfectly flat group delay. The two-times oversampled output data must undergo a final stage of channel filtering in the DSP before being demodulated


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since the transition band region may contain aliased signals, primarily from the adjacent channel.

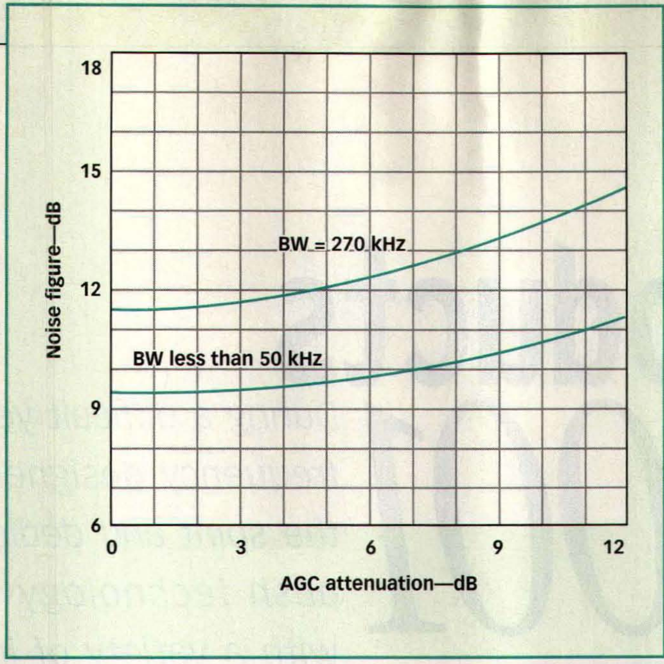
The AD9874 also includes AGC with programmable parameters. The AGC bandwidth spans 50 Hz to 9 kHz and includes adjustable fast attack/slow decay characteristics.

The AD9874 is designed to operate with supply voltages from +2.7 to +3.6 VDC (and up to +5.5 VDC for the charge pumps within the synthesizers) over the extended industrial temperature range of −40 to +85°C. The AD9874 is housed in a space-saving 48LQFP (9.0 × 9.0 × 1.4-mm) package.

An Rx which uses a standard high-power ADC would have a noise figure which is independent of the signal bandwidth, but since the ADC of the AD9874 is optimized for narrowband operation, its noise figure increases as the bandwidth increases. This property is illustrated in **Fig. 4**, which shows that noise figure typically increases by approximately 2 dB as the signal bandwidth goes from 100 to 270 kHz. Below 100 kHz, the noise figure is essentially constant at approximately 10 dB.

It is desirable for noise figure to increase only slowly as AGC attenuation increases, since the Rx's ability to detect a weak signal will then degrade gracefully as interferers increase in strength. This desirable behavior is visible in **Fig. 5**, which shows that as the signal-handling capability of the AD9874's ADC is increased by 12 dB, the noise figure typically increases by only 2 to 3 dB.

Digital Rxs used in GSM cellular base-station equipment must meet some of the most-demanding performance requirements that are found in wireless mobile applications. These Rxs are designed to operate in a hostile radio environment while providing high-quality voice and data services to their mobile handsets. To provide a reliable link, these Rxs require exceptional dynamic range and selectivity to recover target signals that can vary over a 89-dB range in the presence of strong adjacent interferers. Rxs supporting macro cells within the 900-MHz band (i.e., GSM-900) represent the most challenging dynamic-range requirements, while Rxs supporting micro or picocells and Rxs operating in the 1900-MHz band have relaxed interferer and sensitivity specifications as detailed in the European Telecommunications Standards Institute (ETSI) GSM 11.21 specification. In the GSM900 macro-base-transceiver-station (BTS) case, the digital Rx must recover target signals ranging from −15 to −104 dBm while maintaining sufficiently low BER. Since GSM achieves its high capacity by using a time-division-multiple-access (TDMA) scheme whose time slots separated by only 28 μs,



5. The noise figure increases only gradually with AGC due to the high-dynamic-range ADC in the AD9874 (using an IF of 109 MHz and clock frequency of 26 MHz).

the instantaneous dynamic range of the Rx must be high to cope with large slot-to-slot variations in signal strength. The Rx must display good linearity and selectivity to deal with blocker(s) that may fall in adjacent channels with input levels up to −16 dBm at frequency offsets as small as 800 kHz from the target signal.

Figure 6 offers an example of a possible GSM BTS Rx architecture based on this concept. The signal chain consists of a high-linearity RF front end and IF stage followed by two AD9874s operating in parallel. The RF front end consists of a duplexer and preselect filter to pass the GSM RF band of interest. A high-performance LNA isolates the duplexer from the preselect filter while providing sufficient gain to minimize system noise figure. An RF mixer is used to downconvert the entire GSM band to a suitable IF where much of the channel selectivity is accomplished. The 170.6-MHz IF is chosen to avoid any self-induced spurious content from the AD9874. The IF stage consists of two surface-acoustic-wave (SAW) filters isolated by a 15-dB gain stage. The cascaded SAW filter response must provide sufficient out-of-band rejection for the Rx to meets its sensitivity requirements under worst-case blocking-signal conditions. A composite response having 27-, 60-, and 100-dB rejection at frequency offsets of ±0.8, ±1.6, and ±6.5 MHz, respectively, provides enough blocking-signal suppression to ensure that the AD9874 with the lower clip point will not be overdriven by any blocker.

The output of the last SAW filter drives the two AD9874s through a direct signal path and an attenuated signal path. The direct path corresponds to the AD9874 having the lowest clip point and provides the highest Rx sensitivity with a system noise figure of 4.7 dB. The video-graphics array (VGA) of this device is set for maximum attenuation, so its clip point is approximately −17 dBm. Since the conversion gain from the antenna to the AD9874 is 19 dB, the digital output of this path will nominally be selected unless the target signal's power exceeds −36 dBm at the antenna. The attenuated path corresponds to the AD9874 having the highest input-referred clip point and its digital output will be selected under high target signal conditions (i.e., greater than −36 dBm) when the direct path has been overdriven. The input-referred clip point of this path is set to +7 dBm by inserting a 30-dB attenuator and setting the AD9874's VGA to the middle of its 12-dB range. This setting provides a ±6-dB adjustment of the clip point, allowing the clip-point difference to be calibrated to exactly 24 dB so that a simple 4-b shift would compensate for the gain difference. The attenuated path can handle signal levels up to

(Continued on page 126)

Top Products of 2001

During a difficult year, high-frequency designers found the spirit and dedication to push technology forward with a variety of improved and innovative products.

Challenges confronted the microwave/RF industry, and the world in general, during a difficult 2001. Terrorism became a household word after the tragedy of September 11th. The US flirted with recession as more companies were forced to lay off employees to balance their budgets. Despite the hardships, many firms discovered the resolve to develop new and exciting products during 2001, and to push high-frequency, high-speed technologies to new levels.

JACK BROWNE
Publisher/Editor

The list for Top Products of 2001 is based on the combination of technological innovation and practical merit. It combines integrated circuits (ICs), compact subsystems, large systems, and sophisticated test instruments from manufacturers new and old. For example, Agilent Technologies (Santa Rosa, CA), a perennial member of the Top Products list, introduced several products for consideration during 2001, including the powerful E4991A RF impedance/material analyzer, which is capable of making direct impedance measurements through 3 GHz (see April, p. 125). Yet, it was the company's PSG series of performance signal generators that made the list, with outstanding performance from 250 kHz to 40 GHz. The line, which consists of the 250-kHz-to-20-GHz models E8241A and 8251A and the 250-kHz-to-40-GHz models E8244A and E8254A, is characterized by 0.01-Hz frequency resolution, up to +8-dBm output power to 40 GHz, and low phase noise of –110 dBc/Hz offset 20 kHz from a 10-GHz carrier.

Low phase noise was also a key attribute of the model MG8000A frequency synthesizer from Anritsu Co. (Morgan Hill, CA). With 0.1-Hz frequency resolution from 0.1 Hz to 40 GHz and 0.01-dB amplitude resolution from –120 to +17 dBm, the synthesizer combines yttrium-iron-garnet (YIG) source technology with a numerically controlled oscillator for low phase noise with high frequency resolution. The instrument achieves –88-dBc/Hz phase noise offset 1 kHz from a 6-GHz carrier.

When switching speed was the issue, the PTS 6400 frequency synthesizer from Programmed Test Sources (Littleton, MA) provides output signals from 1.000000 to 6399.999999 MHz with 1-Hz frequency resolution. Combining direct analog synthesis with direct digital synthesis (DDS), the PTS 6400 achieves frequency resolution down to 0.1 Hz. The switching speed, which is defined as the time required to settle within 0.1 radian of a new frequency, is a mere 20 µs when switching with 10-MHz resolution. The switching time is even less when switching with smaller digits, typically only 5 µs.

For those requiring modular synthesizers, a line of phase-locked synthesizers from

Top Products of 2001 (in alphabetical order)

Agilent Technologies' PSG series of performance signal generators (May, p. 167)

Analog Devices' AD9874 IF digitizing IC (December cover, p. 106)

Anritsu Co.'s MG8000A frequency synthesizer (January cover, p. 141)

California Eastern Laboratories' UPB1007K GPS Rx chip (July cover, p. 118)

Conexant's CX74017 single-chip GSM transceiver (May, p. 171)

GHz Technologies' 500-W LDMOS FET (August, p. 185)

Micro Lambda's MLSx series frequency synthesizers (March, p. 166)

Palomar's HotRail assembly system (August, p. 203)

Peregrine Semiconductor's CMOS-on-sapphire PLLs (August, p. 196)

Philips Semiconductors' SA2400 2.4-GHz WLAN radio IC (March, p. 157)

Programmed Test Sources 6.2-GHz synthesizer (November cover, p. 92)

Raytheon's 5-GHz WLAN Tondelayo chip set (March, p. 163)

RF Micro Devices' RF3404 CDMA front-end module (May cover, p. 155)

SyChip's miniature GPS module (November, p. 103)

Synergy Microwave's SAW-based Synchronous Optical Network (SONET) oscillators (April cover, p. 115)

Micro Lambda, Inc. (Fremont, CA) features noise floors dropping below –150 dBc/Hz for carrier frequencies through 18 GHz. The YIG-based MLSx series of frequency synthesizers offers 1-Hz frequency resolution, both in narrowband models covering any 2-GHz portion of the 2-to-18-GHz frequency band or in wideband models such as the MLSW synthesizer, which tunes from 2 to 10 GHz. The synthesizers achieve better than +10-dBm output power. The single-sideband (SSB) phase noise is approximately –80 dBc/Hz for a 100-Hz offset from the carrier, dropping to –107 dBc/Hz for a 100-kHz offset from the carrier.

In optical-communications systems, low jitter is essential. A line of voltage-controlled SAW oscillators (VCSOs) from Synergy Microwave Corp. (Paterson, NJ) includes models at 622 and 2488 MHz with better than 100-PPM stability. The phase noise for a 622-MHz VCSO reaches a noise floor of –165 dBc/Hz.

For amplification, a laterally diffused metal-oxide-semiconductor (LDMOS) field-effect transistor (FET) from GHz Technology (Santa Clara, CA) is one of the first LDMOS devices that was developed for traditional high-power pulsed military applications. The Class AB device is capable of 500-W pulsed output power in +26- to +28-VDC 1030/1090-MHz military Identification Friend or Foe (IFF) systems.

Manufacturing these high-performance devices became easier in 2001 thanks to the HotRail Assembly System from Palomar Technologies (San Diego, CA). This automated system provides precision handling of devices, accurate placement of die, and consistent eutectic attachment of devices (placement precision of ±10 to 12 μm). The system handles wafers from 3 to 8 in. (7.62 to 20.32 cm) in diameter and can handle components in 2- and 4-in. (5.08- and 10.16-cm) waffle and gel paks, and in tape-and-reel formats.

Numerous ICs deserve credit during the year for their contributions to high-frequency technology. For example, the AD9874 intermediate-frequency (IF) digitizing IC from Analog Devices (Wilmington, MA) integrates all the function blocks needed for IF-to-digital conversion except the voltage-controlled oscillator (VCO). The IC, which is designed for direct-conversion receivers (Rxs), can capture signal bandwidths as wide as 270 kHz with better than 90-dB dynamic range.

Peregrine Semiconductor (San Diego, CA) has applied its patented Ultra-Thin-Silicon (UTSi) complementary MOS (CMOS)-on-sapphire process to the manufacture of low-noise phase-locked loops (PLLs) based on integer-N and fractional-N approaches. Using CMOS-on-sapphire technology has led to a line of low-power PLLs that are capable of low-phase-noise operation in +3-VDC systems operating up to 3 GHz. The CX74017 direct-conversion transceiver IC from Conexant Systems, Inc. (Newport Beach, CA) cuts the cost of multiband Global System for Mobile Communications (GSM) handsets by eliminating extra IF conversion steps. The CX74017 transceiver is suitable for single-, dual-, or triband GSM handset applications at 900, 1800, and 1900 MHz.

California Eastern Laboratories (CEL; Santa Clara, CA) simplified the design of Global Positioning System (GPS) Rxs with their UPB1007K GPS Rx IC. The IC, which draws only 25-mA current, features an on-board crystal oscillator, multiple mixers, low-noise amplifiers (LNAs), and 40-dB minimum IF downconversion voltage-gain-control range. In modular form, the tiny GPS2020 GPS Rx module from SyChip (Warren, NJ) measures 13 × 15 × 3.75 mm but incorporates an RF Rx, a baseband processor, Flash memory, and a crystal resonator. The GPS2020 module contains several ICs, including a baseband processor, an RF Rx, and memory.

Another module that drew attention was the RF3404 code-division-multiple-access (CDMA) front end from RF Micro Devices (Greensboro, NC). The tiny module, which measures only 8.0 × 8.0 mm, includes an LNA, surface-acoustic-wave (SAW) filter, and mixer. Based on silicon-germanium (SiGe) semiconductor technology, the module requires virtually no off-chip components.

Rounding out the list were two IC solutions for wireless local-area networks (WLANs). The SA2400 radio IC from Philips Semiconductors (Sunnyvale, CA) operates at 2.4 GHz and supports data rates to 11 Mb/s. The IC integrates all the functionality needed for full operation: Rx, transmitter (Tx), synthesizer, VCO, crystal oscillator, and on-chip channel filtering. The four-IC Tondelayo chip set from Raytheon Commercial Electronics (Marlborough, MA) supports WLAN data rates to 54 Mb/s in the 5-GHz Unlicensed National Information Infrastructure (UNII) bands. The set includes a power amplifier (PA)/switch module, a baseband IC, an IF IC, and a frequency converter/LNA IC. MRF

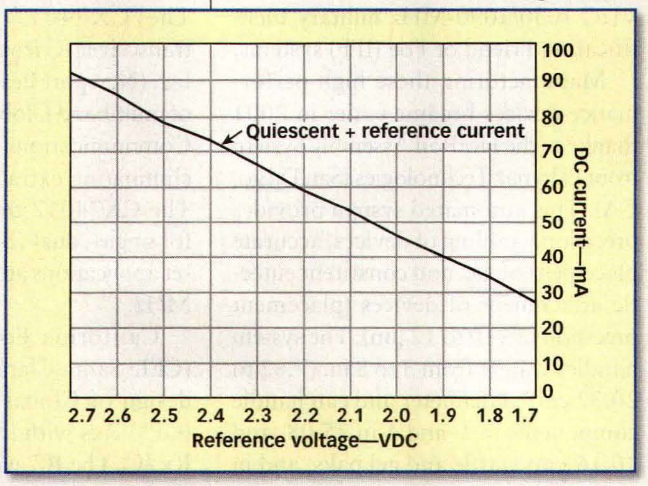
HBT Amplifiers Boast Adaptive Bias Control

Adaptive DC power-management capability using a variable reference voltage on these handset PAs minimizes average current consumption.

JACK BROWNE
Publisher/Editor

Saving power and extending operating life are essential functions for modern cellular and personal-communications-services (PCS) handset designs. Since the power amplifier (PA) in any handset tends to be the largest single consumer of power, it also presents the largest opportunity to save power, an opportunity not lost on the integrated-circuit (IC) designers from the Raytheon's RF Components Division (Andover, MA). The company recently announced the availability of a line of gallium-arsenide (GaAs) heterojunction-bipolar-transistor (HBT) amplifiers employing an "intelligent" bias-control technology known as PowerEdge.(TM)

The new amplifier line includes the models RMPA0951-102 for dual-model cellular handsets (850 to 1910 MHz), RMPA1751-102 for Korean-band PCS code-division multiple access (CDMA) [from 1720 to 1780 MHz], RMPA1951-102 for US-band PCS CDMA and wireless-local-loop (WLL) applications in Korea (1850 to 1910 MHz), and RMPA2051-102 for cdma2000/W-CDMA (1920 to 1980 MHz). The bias-control technology is credited with increasing wireless talk time in these handsets by up to 20 percent for a particular charge or set of batteries. The PowerEdge technology automatically adjusts amplifier bias for maximum efficiency in accordance with input-signal requirements. For example, when relatively close to a base station, less transmit power is needed and thus the PowerEdge circuitry will trim bias current to the PA to a level that is required for



This plot of current consumption versus reference voltage is based on measurements that were made on the model RMPA1951 amplifier.

the appropriate transmit power. Since the PowerEdge approach never uses more power than necessary, it helps to increase overall amplifier power efficiency while maintaining linearity across the operating frequency range.

During transmission, the reference voltage to a PowerEdge amplifier is varied between +1.7 and +2.7 VDC, depending upon the signal strength of a received wireless signal. The lower voltage represents the low-power operating range where most transmissions occur. But even at this voltage, a PowerEdge amplifier can maintain +4-dBm output power with minimum collector current consumption. The resulting benefit is up to a 60-percent reduction in the quiescent and operating currents at that lower limit of the reference-voltage range. As an example of the low quiescent current that is possible with these amplifiers, the current was plotted as a function of reference voltages from +1.7 to +2.7 VDC for a model RMPA1951-102 US PCS amplifier, showing a nearly linear reduction in quiescent current for the reduction in reference voltage **(see figure)**.

As a handset moves away from a base station, and more transmit power is needed, the reference voltage is increased. When higher transmit power levels between +16 and +28 dBm are needed, the reference voltage approaches and reaches typically +2.7 VDC. But even at higher reference voltages, the PowerEdge amplifiers typically achieve better-than-average efficiency, with 20 percent or more reduction in DC power consumption for power levels to +16 dBm, compared to conventional cellular/PCS handset PAs.

The new amplifiers are fabricated with a reliable indium-gallium-phosphide (InGaP) HBT process and are designed for simple operation with a single positive power supply. They are fully matched to 50 Ω at their input and output ports, and supplied in a compact leadless-ceramic-chip-carrier (LCC) encapsulated package measuring only 6.0 × 6.0 × 1.5 mm.

As added benefits to handset designers seeking to save power, the PowerEdge amplifiers offer analog and digital power-management operating modes. The analog mode provides continuously variable current and amplifier gain control (over a range as wide as 10 dB) by continuously varying the reference voltage. The digital mode allows designers to define discrete reference-voltage steps to minimize current and maintain high linearity (ACPR is less than −50 dBc) over specified ranges of output-power levels. The amplifiers also include a power-shutdown mode (when a reference voltage of 0 VDC is

applied) where they draw typically only 2 μA of battery leakage current.

Raytheon RF Components is a leading supplier of components for cellular/PCS handsets and wireless infrastructure equipment. The company is also strongly involved in the development of components for high-data-rate wireless local-area-network (WLAN) systems operating at data rates of 54 Mb/s and beyond **(see sidebar)**. Raytheon RF Components, 362 Lowell St., Andover, MA 01810; (978) 684-8900, Internet: www.raytheon.com/micro.

Enter No. 53 at www.mwrf.com

Raytheon And Systemonic Sign Pact

Developers of 5-GHz high-data-rate wireless local-area networks (WLANs) now have a single source for their integrated circuits (ICs): Systemonic (San Jose, CA). This comes as a result of a multi-faceted agreement between the baseband/digital IC supplier and Raytheon Co.'s Commercial Electronics Business (RCE) this past November. Under the terms of the agreement, Systemonic acquires the products and intellectual property (IP) of RCE's RF Networking group, notably the Tondelayo silicon-germanium (SiGe) RF and intermediate-frequency (IF) chips and a gallium-arsenide (GaAs) power amplifier (PA) for 5-GHz WLANs. These chips, together with Systemonic's HiperSonic baseband processor, form a complete chip-set solution for 54-Mb/s 5-GHz WLAN developers. The programmable baseband processor is suitable to support the multiple protocols of the various versions of the IEEE 802.11 WLAN standard, including the a, b, g, h, and x versions, as well as the European HiperLAN and HiperLAN-2 WLAN standards.

Basically, Systemonic adds RF and IF technology to its product mix while Raytheon receives a focused commercial partner and an equity position in a rapidly growing wireless component developer and supplier. In addition to their equity position in Systemonic, which has offices in San Jose, CA and Dresden, Germany, Raytheon receives a broad license to the current-generation RF technology for use in its military businesses. Additonally, the two companies plan to collaborate on future product developments as channel partners. As part of the agreement, personnel from Raytheon will shift camps and join the smaller, but rapidly growing Systemonic. According to Ruediger Stroh, president and CEO of Systemonic, "We are delighted to welcome a world-class team comprised of RF and protocol firmware designers, as well as marketing and RF operations personnel." Raytheon will supply SiGe wafers and on-going RF IC design services as part of the agreement.

Vector Analyzers Tackle Differential Measurements

These high-performance measurement systems can evaluate the performance of single-ended and balanced components at frequencies through 6 GHz.

High-speed, high-frequency circuit and system designers have employed differential (balanced) architectures for some time to minimize the effects of noise on transmitted signals. Until recently, high-performance test equipment has been geared more toward measurements on single-ended components. But with the introduction of the MS462xD Scorpion Vector Network Measurement Systems (VNMS) from Anritsu Co. (Richardson, TX), engineers can now evaluate single-ended and differential circuits and components through 6 GHz.

DAVID VONDRAN
Product Manager
Anritsu Co., 490 Jarvis Dr., Morgan Hill, CA 95037; (408) 778-2000, (972) 671-1877, Internet: www.anritsu.com.

The new measurement systems **(see figure)** offer balanced/differential measurement ranges of 10 MHz to 3 GHz (the MS4622x system) and 10 MHz to 6 GHz (the MS4623x system) in three- and four-port configurations. The systems can also perform real-time mixed-mode S-parameter measurements, embedding/de-embedding, and impedance transformation calculations. For those with existing systems, upgrades are available to transform MS462xx systems to the new four-port MS462xD balanced/differential configuration or to the three-port MS462xB single-ended/balanced configuration. In either case, these accurate measurement systems reveal the true performance of RF components with typical S-parameter uncertainties of better than 0.05 dB. This type of measurement accuracy further supports the modeling accuracy of computer-aided-engineering (CAE) simulation tools that employ S-parameters in their device and component models.

The MS462xD systems provide accurate vector-network-analyzer measurements on single-ended and differential components through 6 GHz.

Complementing the MS462xD's ability to conduct a wide variety of measurements is its high-end performance. The VNMS achieves a fast measurement speed of 150 Ωs/point, as well as 125-dB dynamic range. It also has up to +10-dBm source power and receiver (Rx) noise as low as −115 dBm **(see table)**.

Since the MS462xD system is based on the Scorpion platform, it can also conduct a number of other measurements to quickly, easily, and thoroughly characterize front-end component performance. For filters, time-domain analysis can simplify tuning. A single connection to an amplifier shows gain compression, harmonics, noise figure,

and intermodulation distortion (IMD). Similarly for mixers, a single connection displays conversion gain, noise figure, IMD, and frequency-translating group delay (FTGD). The MS462xx family ensures thorough and accurate measurements of common front-end components such as antennas, isolators, duplexers, couplers, switches, filters [including surface-acoustic wave (SAW)], and power amplifiers (PAs).

With greater accuracy, innovative flexibility, and these new balanced/differential measurement utilities, the MS462xD system is a powerful new addition to the Scorpion vector-network-measurement systems. It is available in a variety of configurations to satisfy most RF front-end component requirements. **Anritsu Co., 1155 Collins Blvd., Richardson, TX 75081; (800) ANRITSU (800-267-4878), Internet: www.anritsu.com.** MRF

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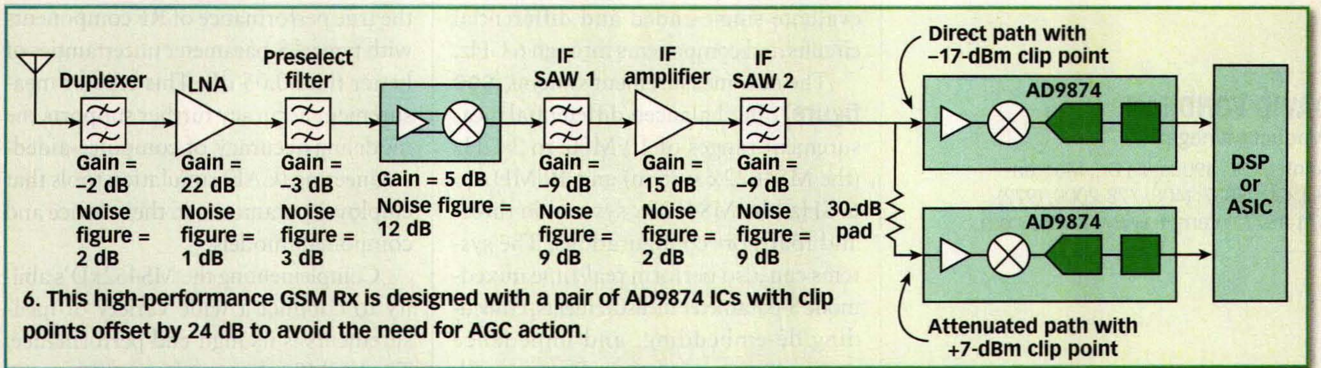
The single-ended/differential analyzers at a glance

TYPICAL SPECIFICATIONS	MS4622X	MS4623X
Frequency range	10 MHz to 3 GHz	10 MHz to 6 GHz
True 3- and 4-port calibrations	Yes	Yes
Passband transmission accuracy	< ±0.05 dB	< ±0.05 dB
Test-port characteristics		
• Corrected directivity	>44 dB	>38 dB
• Corrected port match	>41 dB	>39 dB
• Raw directivity	>23 dB	>23 dB
• Raw port match	>15 dB	>15 dB
Source summary		
• Power range (standard)	+10 to –85 dBm	+7 to –85 dBm
• Level accuracy	±1 dB	±1 dB
• Harmonics	–30 dBc	–30 dBc
Receiver summary		
• Average noise (10-Hz BW)		
• <3 GHz	–115 dBm	–115 dBm
• 3 to 6 GHz		–110 dBm
System dynamic range (Terminated)		
• <3 GHz	125 dB	125 dB
• 3 to 6 GHz		117 dB
High level noise		
• <3 GHz	<0.008-dB RMS	<0.008-dB RMS
• 3 to 6 GHz		<0.018-dB RMS
Measurement speed	150 μs/point	150 μs/point

cover story *(continued)*



6. This high-performance GSM Rx is designed with a pair of AD9874 ICs with clip points offset by 24 dB to avoid the need for AGC action.

(Continued from page 117)

–12 dBm at the antenna before being overdriven. Since the SAW filters provide sufficient blocker suppression, the digital data from this path need only be selected when the target signal exceeds –36 dBm. Although the sensitivity of the Rx with the attenuated path is 20 dB lower than the direct path, the strong target signal ensures a sufficiently high carrier-to-noise ratio (CNR).

Since GSM is based on a TDMA scheme, digital data (or path) selection can occur on a slot-by-slot basis. The AD9874 would be configured to provide serial I and Q data at a frame rate of 541.67 kSamples/s as well as some additional information including a 2-b reset field and a 6-b received-signal-strength-indication (RSSI) field. These two fields contain the information needed to decide whether the direct or the attenuated path should be used for the current time slot.

The AD9874 provides a flexible implementation of the IF-to-bits portion of a superheterodyne Rx. It provides the high performance required for base stations, and the small size and low power consumption needed for handsets. An evaluation board and evaluation software are available. P&A: less than $20 (1000 qty.); 30 days. **Analog Devices, Inc., One Technology Way, P.O. Box 9106, Norwood, MA 02062-9106; (800) 262-5643, (781) 329-4799, FAX: (781) 326-8703, Internet: www.analog.com.** MRF

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looking back

APPROXIMATELY 20 YEARS AGO, Professor Huang Hung-Chia, vice-president of the Shanghai University of Science and Technology (Shanghai, PRC), acknowledged China's need for Western technology in an exclusive interview. He had made a presentation, "Thirty Years of Microwaves in China," earlier that year at the 1982 MTT-S (Dallas, TX).

next month

Microwaves & RF January Editorial Preview

Issue Theme: Test & Measurement

News

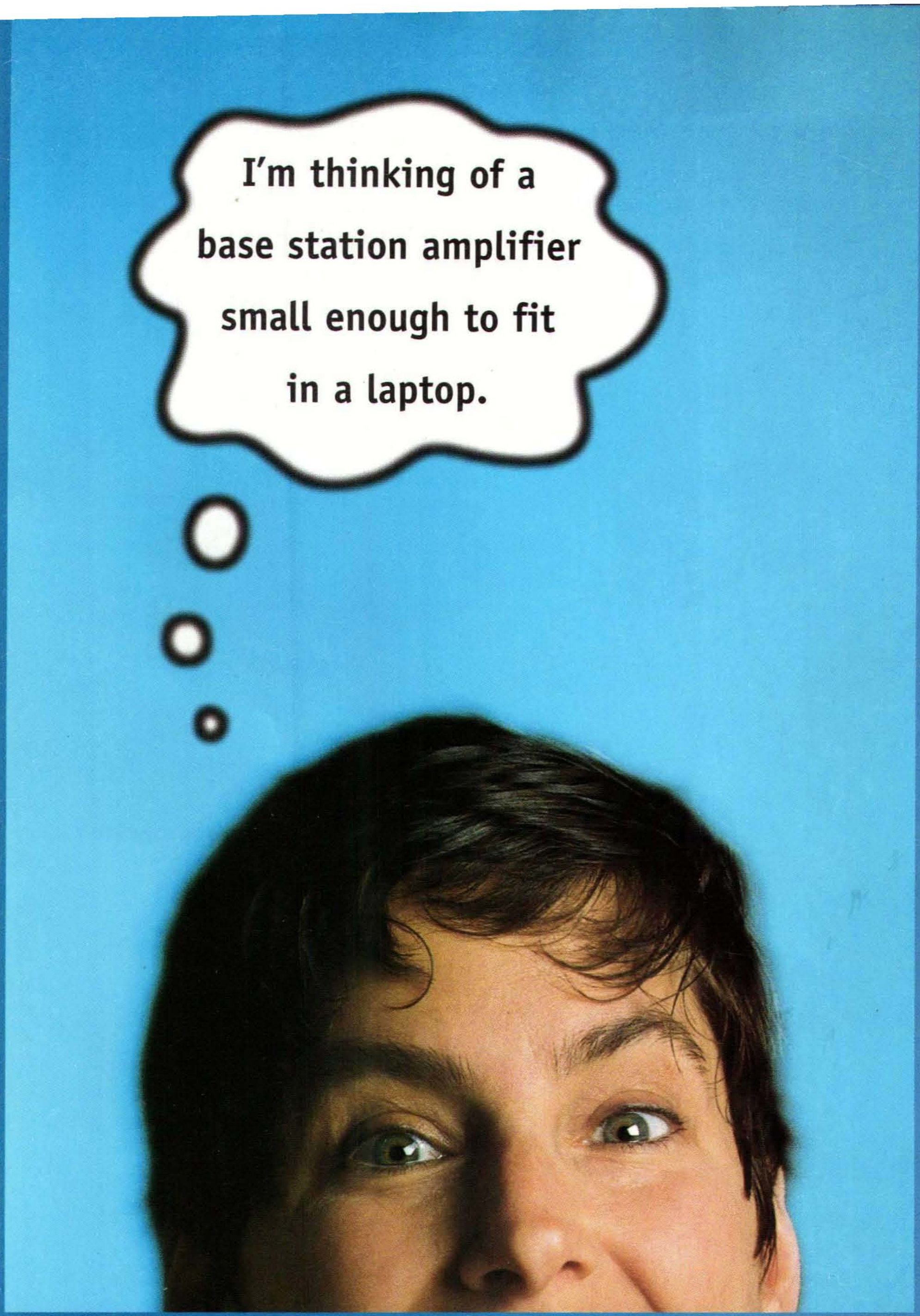
The long-awaited Special Report on TIAs, written by contributor Barry Manz of Manz Communications, will finally appear in the January issue of *Microwaves & RF*. Worth waiting for, this piece will examine the TIA marketplace, from 10-Gb/s CMOS, SiGe, and GaAs devices currently offered by many vendors, to the 40-Gb/s development environment commanded by InP, and explore what manufacturing and test problems remain to be solved. Interviews with several suppliers will project the timing and size of the 40-Gb/s optical-communications marketplace, and the technological capabilities needed to compete.

Design Features

January marks the resumption of a multipart series on short-range radios for wireless data and telemetry applications. Technical articles will also cover biasing techniques for power GaAs FET devices, active and passive RFID technologies, and methods for evaluating packet errors in Bluetooth systems. Additional articles include the second installment of an article series on the design of LNA ICs, the design of amplifiers for NMR systems, and the architecture of a low-noise Rb frequency/time standard.

Product Technology

The January Product Technology section introduces a new approach to noise generators, based on a digital front end that is programmed through numerical codes. Additional product features will examine a software package that helps guide the selection of antenna/cell sites for wireless communications systems, a line of miniature PLLs for cellular and PCS frequencies, a series of high-Q ceramic filters for microwave and millimeter-wave use, a novel 2-to-18-GHz phase modulator, and a software package that combines circuit and EM simulation.



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